

NASA Contractor Report 145269

{NASA-CR-145269} PHOTO SENSOR ARRAY
TECHNOLOGY DEVELOPMENT (Martin Marietta
Corp.) 66 p HC A04/MF A01 CSCI 14E

N78-12393

Unclas

G3/35

53613

Photo Sensor Array Technology Development

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Contract NAS1-14190
December 1977



National
Aeronautics and
Space
Administration

Langley Research Center
Hampton, Virginia 23665



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13 December 1977

To: National Aeronautics and Space Administration
Langley Research Center
Hampton, Virginia 23365
Contract NAS1-14190

Attn: Mr. W. Lane Kelly, Technical Rep.
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Subj: Contract NAS1-14190, Photo Sensor Array (PSA) Technology
Development, Transmittal of Data

Ref: (a) Contract NAS1-14190, Contract Schedule Part V-D
(b) Approval Letter dated Nov. 18, 1977, Signed by
William R. Kivett

Encl: (1) NASA CR-145269, Final Report (10 Copies)

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STANDARD TITLE PAGE

1. Report No. NASA CR 145269	2. Government Accession No.	3. Recipient's Catalog No.	
4. Title and Subtitle Photo Sensor Array Technology Development		5. Report Date December 1977	
		6. Performing Organization Code	
7. Author(s) M. W. Rossman, V. F. Young, J. R. Beall		8. Performing Organization Report No. MCR-77-563	
9. Performing Organization Name and Address Martin Marietta Corporation Denver Division P.O. Box 179, Denver, CO 80215		10. Work Unit No	
		11. Contract or Grant No. NAS1-14190	
12. Sponsoring Agency Name and Address National Aeronautics and Space Administration Langley Research Center Hampton, Virginia 23665		13. Type of Report and Period Covered Technical	
		14. Sponsoring Agency Code	
15. Supplementary Notes N/A			
16. Abstract The development of an improved-capability Photo Sensor Array imager for use in a Viking '75 type facsimile camera is presented. This imager consists of silicon photodiodes and lead sulfide detectors to cover a spectral range from 0.4 to 2.7 microns. An optical design specifying filter configurations and convergence angles is described. Three electronics design approaches; AC-chopped light, DC-dual detector, and DC-single detector, are investigated. Experimental and calculated results are compared whenever possible using breadboard testing and tolerance analysis techniques. Results show that any design used must be forgiving of the relative instability of lead sulfide detectors. Electronics design must be a compromise between gain and drift. A final design using lead sulfide detectors and associated electronics is implemented by fabrication of a hybrid prototype device. Test results of this device show a good agreement with calculated values.			
17. Key Words (Selected by Author(s)) Photo Sensor Lead Sulfide Detector Silicon Photodiode		18. Distribution Statement Unclassified	
19. Security Classif. (of this report) Unclassified	20. Security Classif. (of this page) Unclassified	21. No. of Pages 64	22. Price

PHOTO SENSOR ARRAY TECHNOLOGY DEVELOPMENT

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SUMMARY

The development of an improved-capability Photo Sensor Array imager for use in a Viking '75 type facsimile camera is presented. This imager consists of silicon photodiodes and lead sulfide detectors to cover a spectral range from 0.4 to 2.7 microns. An optical design specifying filter configurations and convergence angles is described. Three electronics design approaches, AC-chopped light, DC-dual detector, and DC-single detector, are investigated. Experimental and calculated results are compared whenever possible using breadboard testing and tolerance analysis techniques. Results show that any design used must be forgiving of the relative instability of lead sulfide detectors. Electronics design must be a compromise between gain and drift. A final design using lead sulfide detectors and associated electronics is implemented by fabrication of a hybrid prototype device. Test results of this device show a good agreement with calculated values.

INTRODUCTION

The Viking '75 lander missions to Mars used facsimile cameras to provide imagery data. These cameras used a Photo Sensor Array consisting of silicon photodiode detectors and associated electronics to image over the silicon response range of 0.4 to 1.0 μm . The characteristics of this system have been described in Reference 1. Several improvements on this system to provide spectrometric capability have been proposed and evaluated. References 2 and 3 investigated the direct-coupled (DC) mode of detector operation using 29 filtered silicon photodiode channels to cover the range from 0.53 to 0.96 μm . Further improvements in the filter spectrometer concept were suggested by Reference 4 which predicted performance and constraints of a system using silicon photodiodes and lead sulfide detectors to cover the spectral range of 0.4 to 2.7 μm . A more detailed examination of this system was performed using computer model simulation (see Reference 5).

The purpose of the present effort is to implement a prototype device which will meet the requirements of Reference 4 in

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general and 5 in particular using hybrid microcircuit technology similar to that used on the Viking PSA. This paper presents the development and investigation of various design concepts proposed to produce a hybrid prototype PSA.

In addition, several tangential evaluations which are essential to the overall PSA development are described. All investigations, analyses, tests, and evaluations performed in the evolutionary development of the working prototype model are also presented.

SYMBOLS

D^*	spectral detectivity, $\text{cm}\sqrt{\text{Hz}}/\text{W}$
f_L	lower 3^{db} cutoff frequency
f_H	upper 3^{db} cutoff frequency
I_{TE}	thermoelectric cooler current, A
T_{TE}	thermoelectric cooler temperature, $^{\circ}\text{C}$
V_C	dark signal correction voltage, V
$\bar{V}_{\text{NO}}$	RMS output noise voltage, μV
V_O	output voltage, V
V_{SUB}	sample and subtract voltage, V
V_{TE}	thermoelectric cooler voltage, V
V^+	detector bias voltage, V

ΔV_{OUT} DC drift output voltage, V

wavelength, μm

DESIGN

This section presents the electronic and optical design of a hybridized Photo Sensor Array (PSA). Design goals, approaches investigated, and approaches implemented are discussed.

Design Goals

The ultimate goal is the development of an improved capability Photo Sensor Array which can operate into the infrared region of the spectrum. This is to be accomplished by the accumulation of knowledge of lead sulfide sensor performance and the design and fabrication of a prototype PSA using hybrid micro-circuit technology.

Specifically, amplification of low level photoconductive detector changes without serious degradation of signal-to-noise ratio is desired. PSA electronics should consume low power, be stable with respect to output drift, and operate at low noise levels over a bandwidth of 140 Hz. Optical design should be implemented with techniques similar to those used on the Viking Photo Sensor Array and should employ standard silicon photodiodes and lead sulfide detectors. Filter holder orientation and incident radiation convergence angles should be designed to produce negligible vignetting when used with an f5.65 optical system. Spectral response up to 2.7 microns is also desired.

Detector and Optics Design

Optical sensing in the visible and infrared regions of the spectrum presently requires the use of two types of detectors. Silicon (Si) photodiodes provide spectral detectivity in the visible region and lead sulfide (PbS) photoconductive sensors, among others, in the infrared region. Since Si photodiode technology is well documented and PbS detector performance in this application is not, investigation of PbS detectors is emphasized. Originally, an Si array of two rows of four cells each and a PbS

array of one row of eight cells was proposed. These configurations are shown in Figure 1. This Figure also defines the proposed PbS optical system including filter holder description, dimensional spacings, and angular calculation. Private communication (2/26/76 letter) and technical discussions with Mr. Bill Davis of ITEK have provided data that indicate for a spacing of 0.020" between filter holder and detector, and 0.020" detector centers there would be no significant vignetting with an F5.6 optics system. It is further indicated that opening up the detector spacing to 0.035" to better accommodate filter sizes and convergence angles would still result in negligible vignetting. This data is based on a 10% off-axis detector location and indicates that approximately 16% vignetting would occur with an F4 system. The Si optical system is the same as used on the Viking PSA.

The PbS array design, then, consists of 0.015 x 0.015" detectors separated by 0.0348" with a 0.020" distance from the filter holder to the detectors for an f5.65 system. It will also include a 0.004" distance from the edge of the filters to the active portion to allow a KERF line for cutting the filters.

Electronics Design

Si photodiode electronics design and performance have been well documented during the Viking program and in literature such as References 2 and 3. The design initially proposed for the hybrid PSA is shown in Figure 2 and is similar to that used on the Viking PSA. It uses transimpedance amplifiers with monolithic FET input operational amplifiers followed by an 8 to 1 analog multiplexer and a post gain stage.

Several methods of amplifying and processing the PbS detector photoconductive changes were proposed and investigated. Design considerations also included hybrid packaging.

An AC-chopped light approach was initially considered because of its advantages in the area of dark signal error correction. That is, if the input radiation is mechanically chopped, alternate signals representing dark and light conditions will be available for processing outside the PSA. The AC approach requires a single row of eight PbS detectors located under the vanes of a tuning fork chopper. The original design using this approach was developed by William R. Patterson, III of Brown University and was transmitted to Martin Marietta in a 23 January

1976 private communication which was also made available to NASA Langley. This letter provides detailed optical and electronic analyses which define and characterize the design in terms of its important parameters (i.e., PbS detectivity, responsivity, noise, and size; optical filter dimensions and convergence angles; chopper dimensions and frequency response; amplifier circuitry, transfer functions, and noise). A summary of recommended primary design characteristics is shown in Table I. This design was to have been implemented with hybrid packaging technology similar to that used on the Viking PSA.

One hybrid packaging scheme which would implement this approach is shown in Figure 3. This Figure shows placement of the chopper in the center of the hybrid package (with Si photodiodes exposed) flanked by two hybrid substrates containing the detector amplification circuitry. Chopper requirements, coordinated with Bulova, specify a chopper which would operate at 800 Hz and have stainless steel blades which would be coated black for low light reflectance from 1.0 to 2.7 microns. PbS amplification circuitry is shown in Figure 4 and employs 6 db/octave bandpass preamplifier stages, tuned second-stage amplifiers and chopper drive control demodulator sections. One disadvantage of this system is the requirement to process the output signal from AC back to DC outside the PSA hybrid. A further disadvantage is the increase in mechanical complexity which may cause a decrease in overall system reliability.

The next approach considered was a DC-dual detector system. This system was designed to eliminate the need for dark signal correction by external signal processing circuitry and to provide drift compensation for the inherent instability of PbS detectors. Physically, this approach would eliminate the chopper shown previously in Figure 3 and would require two rows of eight PbS detectors. One row is to be kept dark to provide a reference signal which is used for internal dark signal correction. The other row (sensing cells) is located under the filter holder as was shown previously in Figure 1. An important optical consideration in this system is the row-to-row detector spacing which must be large enough to preclude stray light from affecting the dark (reference) cells. The circuitry required (see Figure 5) is similar in some aspects to a bridge configuration in that the output from the reference (dark signal) amplifier is "balanced" against the sense cell output at an amplifier summing node. This system embraces the features and advantages of the AC design, but also provides low frequency, $1/F$, noise rejection since the capacitor in the reference amplifier feed-

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back can be made relatively large (determines low frequency cut-off). The dual detector system also eliminates the need for external correction, but its success depends heavily on the drift characteristics of the detectors. Documentation of this design including definition and characterization of its most important parameters is provided by the tolerance analysis shown in the Mathematical Analyses section of this report.

The final approach considered and eventually implemented in hybrid form was a DC-single detector system which evolved of necessity due to certain problems encountered during investigation of PbS detectors and breadboard testing. The physical configuration of this approach is similar to the DC-dual detector with the PbS reference cells removed. However, the addition of a thermoelectric cooler provides for the thermal stability of the detector array. It is required to maintain the detector array at $25 \pm 1^\circ\text{C}$ for ambient temperatures between approximately -48 and $+63^\circ\text{C}$. In the amplifier circuitry the reference amplifier is replaced by a DC correction voltage. Figure 6 shows the circuitry required to produce this system. Documentation of this design including definition and characterization of its most important parameters is provided by the tolerance analysis shown in the Mathematical Analyses section of this report.

Due to the nature of this program (technology development), design approaches were investigated and evaluated until they were found to be impractical or until another approach was shown to be superior. Further, the depth of investigation necessarily changed as new facts were uncovered. Therefore, as certain areas were more intensely studied, other areas which were more well defined were eliminated from study. Such progressive narrowing down of perspective occurred on this development program. The AC approach was initially investigated until it was found to be incompatible with existing Viking camera design requirements. While the DC-dual detector approach met these requirements, it became relatively impractical due to lack of TCR matching between reference and sense PbS detectors. Finally, it was determined that a deeper understanding of the behavior of PbS detectors was essential to the development of a PSA design. Hence, the compromise design of the DC-single detector approach emerged as the best way to provide a hybrid test vehicle where PbS characteristics could be studied to the exclusion of Si detectors, multiplexing techniques, and post-amplification techniques. Further, optical implementation and testing was also eliminated due to the heavy emphasis placed on detector technology development and multiple design investigation.

MATHEMATICAL ANALYSES

Mathematical analyses were performed to support PSA design efforts. This section describes tolerance analyses of the DC-dual detector and the DC-single detector approaches and a thermal analysis of thermoelectric cooler requirements.

Tolerance Analyses

These analyses consist of calculations of DC drift, AC transfer function, and noise characteristics based on circuit component values and tolerances. They predict breadboard and hybrid circuit performance and identify worst-case conditions. DC drift determination is accomplished by calculating output sensitivity to detector, offset voltage, bias current, and circuit resistor drifts. The AC transfer function in the complex frequency domain is obtained from a gain model which includes first pole and feedback time constants. An output RMS noise equation is obtained from a noise model constructed from detector, Johnson resistor, bias current shot, and amplifier noise sources. Summarized results are shown for both DC approaches in Table II. It should be noted when comparing the two approaches that the dual detector analysis assumed a gain of 15 while the single detector assumed a gain of 30. This is because detectors with dark resistances of 1 m Ω were used in the dual detector breadboard and 500 K Ω in the single detector hybrid prototype. Complete analyses may be seen in Appendices A and B, for the dual detector and single detector, respectively.

Thermal Analysis

A thermal analysis was performed to define thermoelectric cooler/heater requirements. It was assumed that the detector array was to be maintained at 25°C for ambient temperatures between -48 and + 63°C. For this analysis 25°C was chosen as a worst-case condition in terms of the heater/cooler operation. It is recognized in actual application that lower temperature operation is recommended for a more efficient detector performance. The detailed analysis is shown in Appendix C. Results show that heating the detector from a low ambient condition is the worst case and that the maximum power required is 0.469 watts.

EXPERIMENTAL EVALUATION

This section presents the results of PbS detector supplier and user survey, PbS detector testing, DC-dual detector bread-board testing, and DC-single detector hybrid prototype testing.

PbS Detectors

Early tolerance analysis results showed that PbS detector characteristics would dominate PSA performance parameters such as drift and noise to such an extent that a thorough investigation of these detectors was warranted.

Published PbS detector technical data were reviewed to obtain a better understanding of device physics and performance. The PbS detector is a polar semiconductor with an N-type composition in the unoxidized form.

The detector film is a polycrystalline structure which is typically 1.0 micron thick. The films are produced by one of two methods, vacuum evaporation or chemical deposition. Of these two, chemical deposition is the most widely used process. Following film deposition they must be sensitized to optimize their photoconductive response. For a vacuum evaporated PbS film sensitizing is accomplished by evaporating the film in a low pressure oxygen atmosphere. For a chemically deposited PbS film there are two methods for sensitization. One is by baking film at a high temperature in an oxygen atmosphere. The second is by introduction of a chemical oxidizing agent in the chemical bath. Following deposition the film is vacuum baked to optimize the photoconductive response. This second method provides a better film consistency and the process is more easily controlled. The crystallite size in an unoxidized film is in the order of 0.5 microns. Oxidation of the film decreases the crystallite size to about 0.1 microns. Uniformity of crystallite size is an important factor in obtaining uniform photoresponse across the detector film. The oxidation of the PbS film changes the surface from an N-type to a P-type semiconductor. The P-type thickness has been postulated to be in the area of 300 Å.

Being a polycrystalline structure, the PbS film contains a multiplicity of energy band gaps. Therefore, photoconductive response and carrier conduction in the PbS are complex mechanisms and have received a great deal of study and analysis. Two

theoretical mechanisms of photoconductive response have been postulated. One is that photons generate free carriers which increase the conductivity of the detector. The second is that photons generate free carriers which are trapped in the film. The trapped carriers reduce the potential barriers, thereby increasing the effective carrier mobility which in effect increases the detector conductivity. For PbS detectors the applicable mechanism appears to be the first postulation of free carrier generation.

Also, as a polycrystalline film, the physical properties exhibit extremely wide variances between film deposition lots. Therefore, to obtain optimum matching of detector characteristics the detectors must originate from the same deposition process lot.

A survey was made of PbS suppliers and users. Three suppliers were contacted regarding the application of PbS detectors in DC circuit applications. The suppliers contacted were Infra-red Industries in Waltham, Massachusetts, Santa Barbara Research Center in Goleta, California, and Opto Electronics in Petaluma, California. The dark resistance TCR tracking typically runs from 5% to 20% for off-the-shelf detectors. The general consensus was that a 1% TCR tracking between matched detectors would be difficult to meet. A major concern was the extreme sensitivity of the detector films to moisture. This sensitivity can be reduced by using a coating to seal the film surface. The PbS detector resistances range from 0.5 to 5.0 megohms with typical being 1-2 megohms. There was no data available regarding aging effects on TCR tracking. The maximum temperature rating for PbS detectors is + 70°C. This severely limits the possible utilization of accelerated testing to evaluate aging effects and operating life performance. Very little application information was available regarding DC circuit applications. The majority of detector applications is involved with classified military programs and this information is not available. There are three detector parameters which are very sensitive to temperature changes, dark resistance, D^* , and responsivity. These parameters exhibit a negative temperature coefficient of typically 3-4%/°C. D^* and responsivity improve with decreasing temperature. These three parameters can be degraded by flashing the detector. Flashing is the exposure to ultraviolet or strong visible light. Dark resistance decreases due to flashing. The degree of degradation cannot presently be predicted. It is dependent on the intensity, wavelength and duration of exposure and this reaction has not been characterized. Generally, for

PbS detectors, if the detector is allowed to remain in a dark condition it will recover in about one day or at a temperature of 50-60°C in about one hour.

Four users of infrared detectors were contacted. They were Martin Marietta/Orlando, Night Vision Lab at Fort Belvoir, Virginia, Hughes Aircraft, and JPL. Martin/Orlando and Night Vision Lab had no experience with PbS detectors. Hughes Aircraft has applications utilizing PbS detectors but they are on classified contracts and they would not discuss them. JPL used PbS detectors on the Mars Orbiter (Mars Atmospheric Water Detection System, MAWD). These detectors were radiation cooled to maintain temperature stability. Difficulty was experienced in identifying or defining the best quality and performance that could be obtained from the supplier. This information was classified by the programs for which they were developed, so JPL elected to screen the rejects from these military programs. They felt that they were able to obtain better detectors using this method than what could be obtained from off-the-shelf product. From this screen they selected the best detector for each application. This necessitated the selection of detector bias voltage and compensation resistor for each detector in the system. They were not aware of stabilization or screening tests which could be utilized in the evaluation of detector reliability.

A literature search to identify recommended visual inspection and electrical screening criteria has been unsuccessful. Very little information was available from the suppliers as well.

Fifteen detectors which were deposited on three different substrate materials were obtained for evaluation. The substrates were polished Al_2O_3 , sapphire and quartz. This evaluation primarily addressed the dark resistance parameter with respect to temperature. The detector elements were tested at five different temperatures between 20 and 60°C using various time increments and temperature sequence. The test circuit utilized was equivalent to that used by the supplier. The detector bias is applied to the detector through a 1 megohm series load resistor. The detector resistance is extrapolated from the detector voltage drop and circuit current. The detector bias voltage used was as specified by the detector manufacturer. The detectors were covered to preclude exposure to direct light or IR from the test chamber heater.

A summary of these data showed:

- 1) Thermal stabilization following the application of detector bias was greater than two hours for an initial ambient temperature of 20°C.
- 2) The delta temperature generated by the detector film power dissipation ranged from 1.0 to 4.2°C.
- 3) Extrapolated TCR's using least squares linear regression ranged from 9.5K to 17.2K ohms per °C. The mean for all detectors measured was 12.3K ohms per °C. This is equivalent to an average of 1.4%/°C over the range of 20-60°C.
- 4) The TCR slopes ranged from -7.0×10^{-5} to -8.2×10^{-5} for polished Al_2O_3 , -6.9×10^{-5} to -8.7×10^{-5} for sapphire and from -1.0×10^{-4} to -1.1×10^{-4} for quartz. This showed the tracking quality for detectors deposited on quartz to be far better.

DC-Dual Detector Breadboard

The DC-dual detector design was breadboarded using discrete components and detectors to provide an experimental comparison against design goals and tolerance analysis values. Two amplifier channels were constructed and two Optoelectronics OE-20 discrete packaged detectors were used. Table III shows the performance characteristics of the detectors as tested by the supplier. Throughout the testing and evaluation, the two amplifier channels produced very similar results. Because of this close similarity, results in the following paragraphs will be shown for one channel to avoid unnecessarily cluttering the visual presentations.

DC drift tests were performed to evaluate stability. Breadboard and detectors were maintained at a room ambient of $22 \pm 3^\circ\text{C}$ under dark conditions. All power supplies and test equipment were monitored periodically to ensure accuracy of measurement. The results of this test are plotted in Figure 7 and show a relatively large amount of drift with a very long stabilization time constant (apparently due to power dissipation heating effects). Calculations using this data show a 2.3% relatively sensor drift (resistance change) over the 1-week test

period. This compares rather poorly with the 1% drift design goal assumed in the tolerance analysis. Measurements of drift due to the electronics alone (with detectors simulated by film resistors) show the amplifier circuitry contributes less than 0.1% drift to the relative sensor drift calculated above. Overall output drift (4.90 volts) is also seen to exceed the 3.61 volts calculated in the tolerance analysis. Repeated test runs show little hysteresis effects. When testing was repeated at + 45°C, similar data were gathered.

Frequency response measurement was performed to determine gain characteristics and break frequencies. Detectors were simulated with 1 MΩ film resistors and a sinusoidal signal was applied. The breadboard was tested at room ambient temperature by measuring input and output peak and RMS voltages at discrete frequency intervals. The frequency response plot is shown in Figure 8 and compares very favorably with tolerance analysis calculations. Upper and lower break frequencies can be seen to be approximately 0.09 Hz and 125 Hz, respectively, which compares with a design bandwidth of 0.1 to 140 Hz. The upper break frequency deviation is apparently due to stray wiring capacitance on the breadboard. Results also show that peak gains obtained (≈ 25 db) are also consistent with tolerance analysis values (gain $\approx 15 \approx 24$ db).

Noise measurements were performed on the breadboard with detectors simulated by film resistors to determine noise contribution due to amplifier circuitry alone. Wave analyzer techniques were used to determine frequency components of the noise. The data were reduced using calculations based on the noise bandwidth of the wave analyzer. Figure 9 shows the noise characteristics of the amplifier circuitry between 20 and 160 Hz. Extrapolations from this curve and first order approximation calculations of average noise verify the tolerance analysis assumption that the detectors are the dominant contributors to noise in the system. Extrapolation of the curve back to nearly 0 Hz and calculating the average noise over 0 to 140 Hz yields a noise voltage of approximately $16 \mu\text{V}/\sqrt{\text{Hz}}$. Eliminating the detector noise contribution from tolerance analysis calculations over 0.1 to 140 Hz yields $64 \mu\text{V}/\sqrt{\text{Hz}}$. This data indicates that the amplifier circuitry performs much better than the tolerance analysis indicates. In fact, the tolerance analysis appears to be an extreme worst case.

DC-Single Detector Prototype Hybrid

The DC-single detector design was used to fabricate and package a prototype hybrid PSA which is illustrated in Figure 10. The PSA hybrid contains seven PbS preamplifiers on an alumina substrate which is constructed with screen printed and furnace fired thick film gold conductors, thick film chip resistors, ceramic chip capacitors, and chip integrated circuits. The PSA also contains an 8-cell PbS array custom deposited on a fused quartz substrate by Optoelectronics. A "sandwich" configuration method of construction is used with a thin cover of quartz bonded on top of the detectors with Araldite 6060 epoxy. This provides physical protection for the detectors and the detector overcoating (which acts as a moisture barrier). Table IV summarizes the array characteristics as tested by the supplier. This array is mounted on top of two Marlow Industries 1020 thermoelectric coolers with thermally conductive epoxy. The coolers are electrically connected in parallel with a common wire grounded such that operation is always in the cooling mode.

Normal fabrication and assembly techniques and materials were used to produce this hybrid. Aluminum ultrasonic and gold thermocompression wire bonding maintains a nearly monometallic system while epoxy chip, substrate, and cooler mounting provide lower temperature stressing during assembly than would eutectic bonding. Hybrid circuit packaging technology similar to that used on the Viking PSA was employed. The aluminum package used is black anodized to minimize light reflectance and is hermetically sealable. Early hermeticity tests using a plastic O-ring produced leak rates less than 1×10^{-7} Atm. cc/sec.

Several precautions were taken during hybrid assembly to prevent detector flash damage due to exposure to ultraviolet or strong visible light. During non-assembly storage detectors were kept in "black boxes" which were opened only under low, indirect lighting. Assembly stations were shrouded to exclude fluorescent lighting and microscope lights were reduced to minimal levels. Supplier data recommends that light exposure levels be less than 15-foot candles. After hybrid assembly was complete, an attempt was made to actively trim the 500 K Ω resistor at the V_C input while monitoring dark level output voltage. Trimming was accomplished with a laser, but positioning of the resistor under the laser beam was done with a microscope light several times brighter than 15-foot candles. Exposure to this light was approximately two minutes. During trimming it became

obvious that some kind of degradation had taken place. Measurements taken before and after the exposure indicated that detector resistance had been reduced from its original value by a factor of 0.94 on the average. Further measurements showed the detectors recovered to their original values approximately four hours after removal to "black box," room ambient storage.

The hybrid PSA prototype was tested to provide drift, frequency response, noise, responsivity, and thermoelectric cooler performance data. This data may be compared against tolerance analysis calculations made previously.

DC drift tests were performed under dark conditions with the detector temperature controlled to $20 \pm .25^\circ\text{C}$. Correction voltages were applied initially to "null" the circuit (set output voltage to zero). The results of a typical and the worst-case channel are plotted in Figure 11. This curve shows an apparently large drift with a relatively short stabilization time constant. Since the circuit gain is larger than that of the dual detector approach, the output voltage drift is deceptive. Obviously, such a high circuit gain is undesirable due to dynamic range considerations. However, when the output drift is referred back to the detector, the detector drift is seen to be 2.3% typically and 2.7% worst case which again compares unfavorably with the 1% drift design goal assumed in the tolerance analysis. Overall output drift (7.20 volts typical and 8.00 volts worst case) is also seen to exceed the 5.94 volts calculated in the tolerance analysis.

Frequency response measurements are plotted in Figure 12 and compare very favorably with tolerance analysis calculations. The upper break frequency calculated at 156 Hz from the tolerance analysis is seen to be approximately 160 Hz in the hybrid prototype. Peak gain (28.9 db) is also very consistent with tolerance analysis values (gain $\cong 30 \cong 29.5$ db).

Noise measurements were performed on the hybrid prototype for comparison with tolerance analysis worst-case values. Figure 13 shows the hybrid PSA noise characteristics between 10 and 160 Hz. Extrapolating the curve back to nearly 0 Hz and calculating the average noise over 0 to 140 Hz yields a noise voltage of approximately $353 \mu\text{V}/\sqrt{\text{Hz}}$. This can be compared directly with the $455 \mu\text{V}/\sqrt{\text{Hz}}$ value obtained in the tolerance analysis. Again it is found that the tolerance analysis appears to be an extreme worst case condition.

Detector spectral response testing was accomplished by exposing detectors to filtered (1.8, 2.25, 2.5, 2.7, 3.0, and 3.5 μm) radiation from a known blackbody source. The blackbody was run at a temperature of 858 $^{\circ}\text{K}$ with a chopping frequency of 600 Hz. Detectors were maintained at an operating temperature of +20 $^{\circ}\text{C}$ with output measurements made alternately at dark and irradiated conditions. Blackbody output power calculations based on physical parameters (detector area, source-to-detector distance, etc.), known blackbody response, and known filter transmissibility were combined with detector output measurements. The resulting NEP and D^* characterizations are shown in Figure 14. Spectral response data is compatible with vendor data within the limitations of the test. Since the entire hybrid was "flooded" with blackbody radiation, detectors were exposed to relatively large background radiation and amplifier chip integrated circuits were also exposed to radiation. These conditions could have contributed to the uncharacteristically high response below 2.5 μm which should have been the approximate peak. This problem is easily corrected by performing the test on totally packaged hybrid (i.e., with filter holder, filters, and windowed lid in place) and using a blackbody with aperture control.

Thermoelectric cooler performance was evaluated by reducing the detector temperature, as measured by thermocouple, from room ambient conditions and noting electrical characteristics. Figure 15 shows the current change required to produce various degrees of cooling. Figure 16 shows current-voltage characteristics. Remembering that two coolers are being used in parallel, these figures can be extrapolated to show single cooler performance. Such an exercise shows that the coolers, as installed, perform significantly less well than supplier specifications indicate they should. It should be noted here that installation of the coolers in the hybrid PSA requires epoxy attach processing with a 130 $^{\circ}\text{C}$, one-hour cure. Coolers are not rated for operation over 80 $^{\circ}\text{C}$, but maximum storage temperature specification is unknown. Recommended mounting temperature is 96 $^{\circ}\text{C}$.

CONCLUDING REMARKS

This paper has presented several approaches to the development of an improved-capability photo sensor array for use in a filter spectrometer facsimile camera. Overall PSA design including optics, silicon photodiode technology, and silicon

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sensor amplifier circuitry were described. The characteristics of lead sulfide detectors and three design approaches to process PbS photoconductive changes were investigated. A hybrid prototype PSA using a PbS detector array was fabricated and evaluated.

Overall PSA optical design using filter holder and windowed package lid has been well documented and provides a sound method of light processing. Silicon photodiode technology and amplifier circuitry are similarly well investigated providing a well established method of imaging. However, it is noteworthy that imaging technology using charge-coupled devices, CCD's, is becoming increasingly important as it is steadily developed for use in practical applications. In fact, presently estimated size, weight, resolution, and other technical advantages over conventional imaging techniques would appear to make CCD imaging more desirable than silicon photodiode technology.

Lead sulfide detector evaluations produced evidence that inherent thermal instability as shown in relatively large thermal coefficients of resistance requires the accurate controlling of detector temperature. This is particularly disadvantageous since control accomplished by conduction or radiant heating/cooling implies an increase in PSA complexity due to the need for techniques of temperature measurement (thermocouple) and management (thermoelectric device, etc.). Additionally, the instability in DC drift creates problems in processing and amplification circuitry by restricting the useful gain due to limitation in dynamic range.

The three PbS amplifier circuitry designs performed well when compared to predicted and calculated parameters. However, each had at least one objectional characteristic. The AC-chopped light approach appears very promising due to its ability to provide dark signal and drift correction, but it requires more processing circuitry external to the PSA to convert AC back to DC signals. Furthermore, the added mechanical complexity and associated decrease in reliability is undesirable. The DC-dual detector is a sound approach to self-contained internal dark signal and drift correction, but is restricted in usefulness by PbS detector instability and lack of tracking. The DC-single detector approach provides a vehicle with which to evaluate detector and amplifier circuitry performance, but is undesirable because of the external correction signals required for each channel.

Production of a hybrid prototype PSA using PbS circuitry only was shown to be feasible. Standard hybrid fabrication and

assembly methods may be used with some precautionary techniques necessary. Important considerations are the avoidance of flash damage to PbS detectors and assuring that thermal exposure during assembly does not exceed individual component, (detector array and thermoelectric cooler in particular) maximum temperature.

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TABLE I. AC-CHOPPED LIGHT DESIGN
CHARACTERISTICS

Type of Detector:	Room temperature lead sulfide.
Number of Channels:	Two channels active at all times with separate bias connection for each one.
Size of Detectors:	.015" square on .035" centers.
Bandwidth:	140 Hz.
Preamplifier Circuit:	See Figure 4.
Chopper:	Bulova type L2-C with frequency 800 Hz and bright finish blades mounted as closely as possible to the base plane. The blades are .310" long with a rest separation of .025" nominal.
Case Material:	Nickel plated brass.
Window:	Schott RG-780 glass.
Case Seal:	Silicone rubber, if needed.

TABLE II. DC PSA ELECTRONICS TOLERANCE ANALYSIS CALCULATIONS

DC DRIFT ΔV_{OUT}	AC 3 ^{db} BREAK FREQUENCIES		OUTPUT RMS NOISE, $\bar{V}_{NO}$
	LOWER, f_L	UPPER, f_H	
	DUAL DETECTOR SYSTEM		
3.61 Volts	0.1 Hz	140 Hz	375 Microvolts
	SINGLE DETECTOR SYSTEM		
5.94 Volts	N/A	156 Hz	455 Microvolts

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TABLE III. SUPPLIER TEST DATA ON OE-20 PbS DISCRETE DETECTORS

ELEMENT SERIAL NUMBER	DETECTOR DARK RESISTANCE	DETECTOR BIAS (VOLTS)	SIGNAL (μ VOLTS)	NOISE (μ VOLTS)	SIGNAL TO NOISE RATIO	RESPONSIVITY (λ pk, $\frac{600}{\text{VOLTS/WATT}}$)	D* (λ pk, $\frac{600}{\text{CM Hz}^{1/2} \text{ W}^{-1}}$)
	AMBIENT (M Ω)						
1	.722	12	125	1.5	83	3.39×10^6	1.79×10^{11}
3	.915	12	165	2.0	83	4.48×10^6	1.78×10^{11}

TEST CONDITIONS:

Blackbody Temperature 500 Degrees Kelvin
 Flux Density 6.25×10^{-6} Watts/CM²
 Chopping Frequency 600 Hertz
 Noise Bandpass 10 Hertz
 Operating Temperature 298 Degrees Kelvin
 Heat Sink Temperature N/A Degrees Kelvin
 Detector Element Size .025 x .025 CM
 Area 6.25×10^{-4} CM²

$$\frac{D^* (\lambda \text{ pk})}{D^* (\text{BB})} = \underline{106}$$

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TEST CONDITIONS:

Blackbody Temperature	<u>500</u>	Degrees Kelvin	Heat Sink Temperature	<u>N/A</u>	Degrees Kelvin
Flux Density	<u>6.25×10^{-6}</u>	Watts/CM ²	Detector Element Size	<u>.0381 x .0381</u>	CM
Chopping Frequency	<u>600</u>	Hertz	Area	<u>1.45×10^{-3}</u>	CM ²
Noise Bandpass	<u>10</u>	Hertz	D* (λ pk)	<u>= 106</u>	
Operating Temperature	<u>298</u>	Degrees Kelvin	D* (BB)		

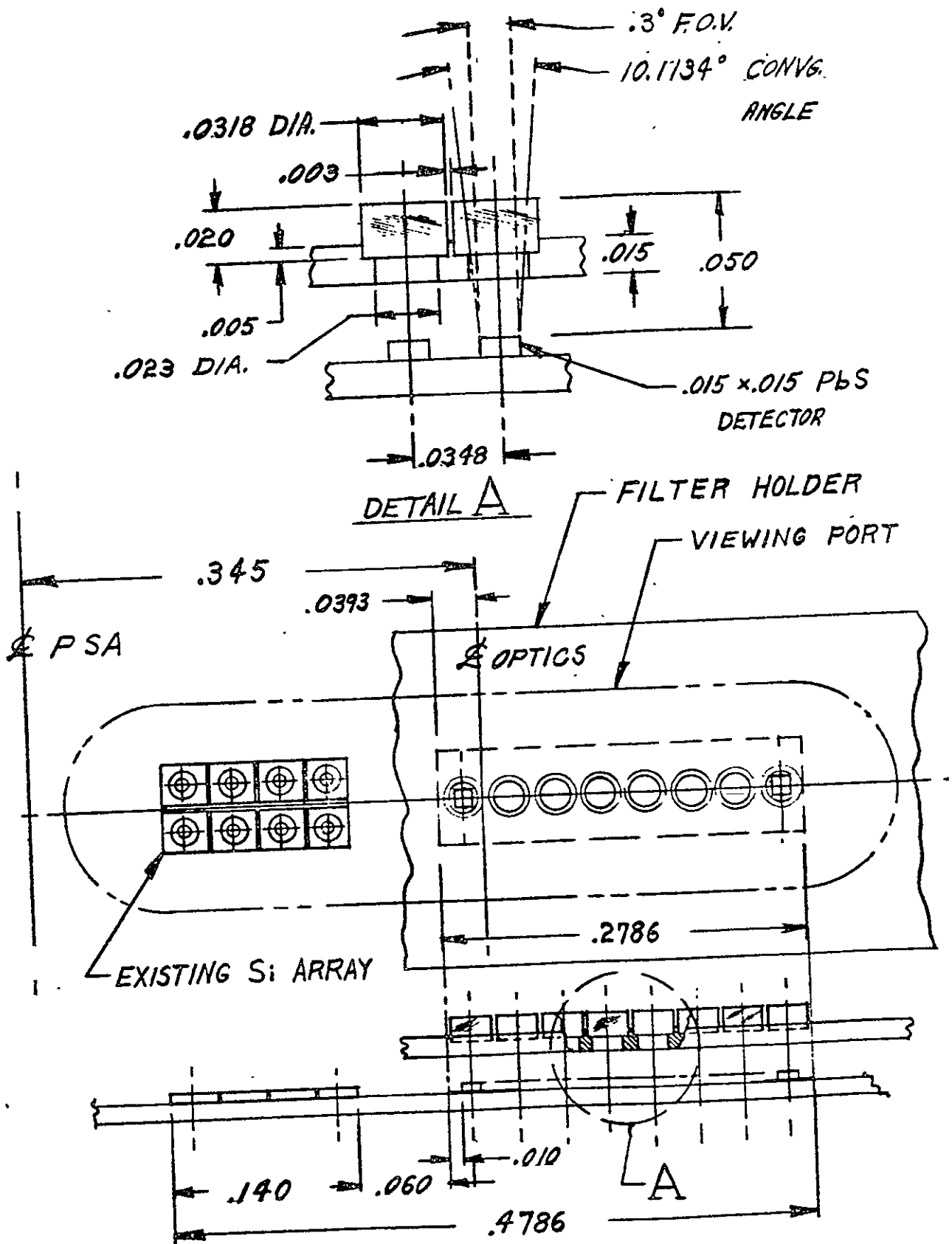


Figure 1. Detector and Optical Configuration

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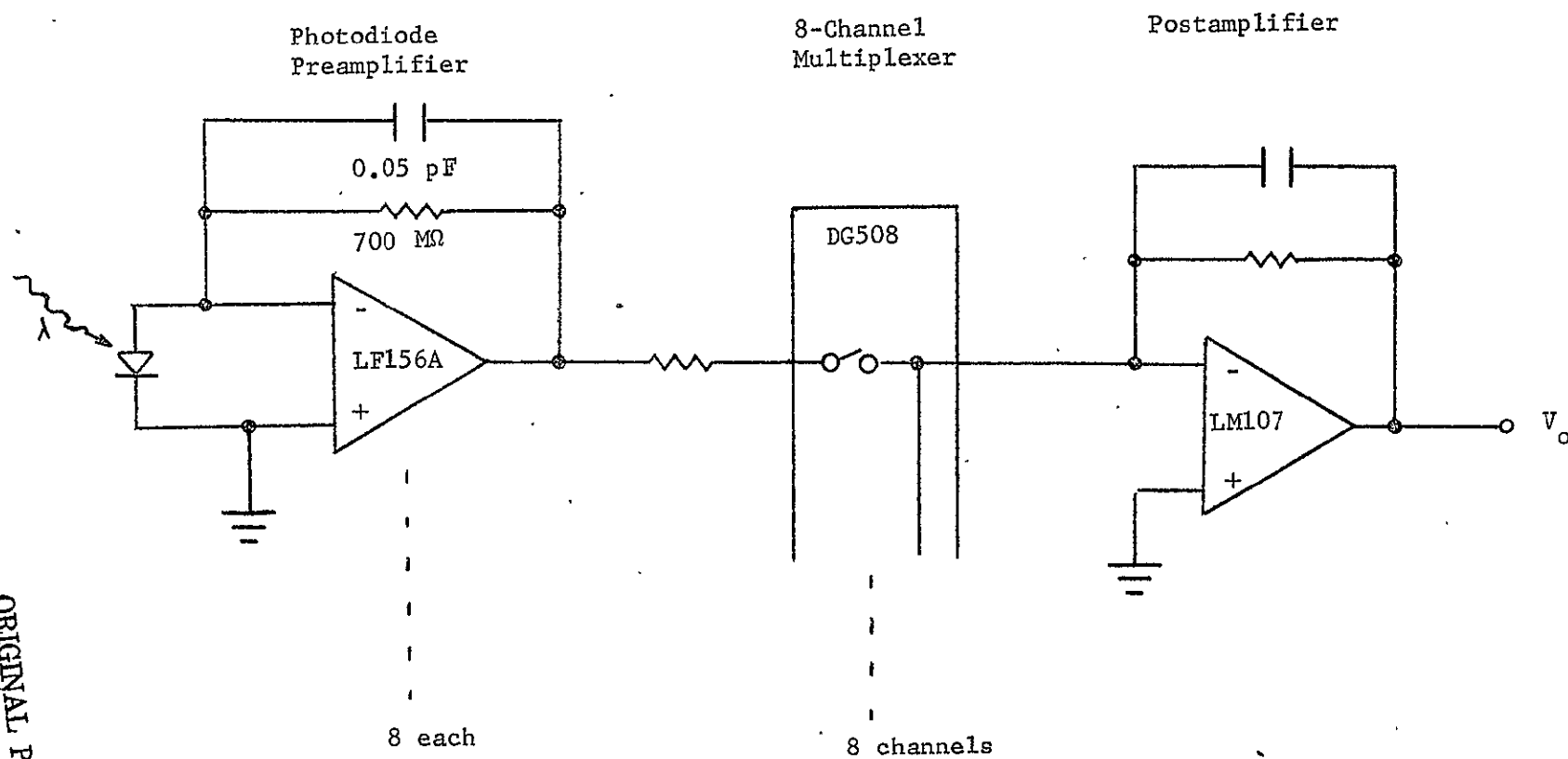


Figure 2. Silicon Photodiode Amplifier Circuitry

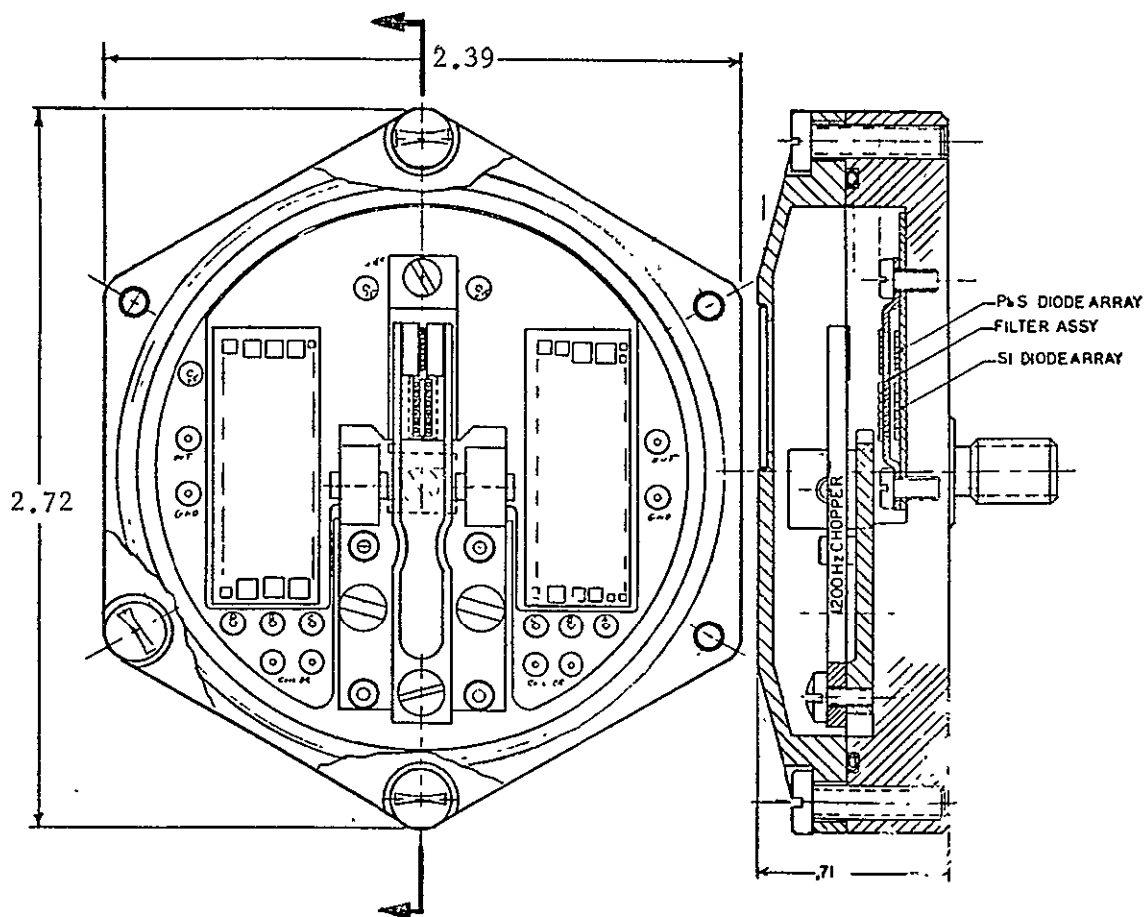


Figure 3. Physical Configuration of Proposed AC-Chopped Light PSA Hybrid

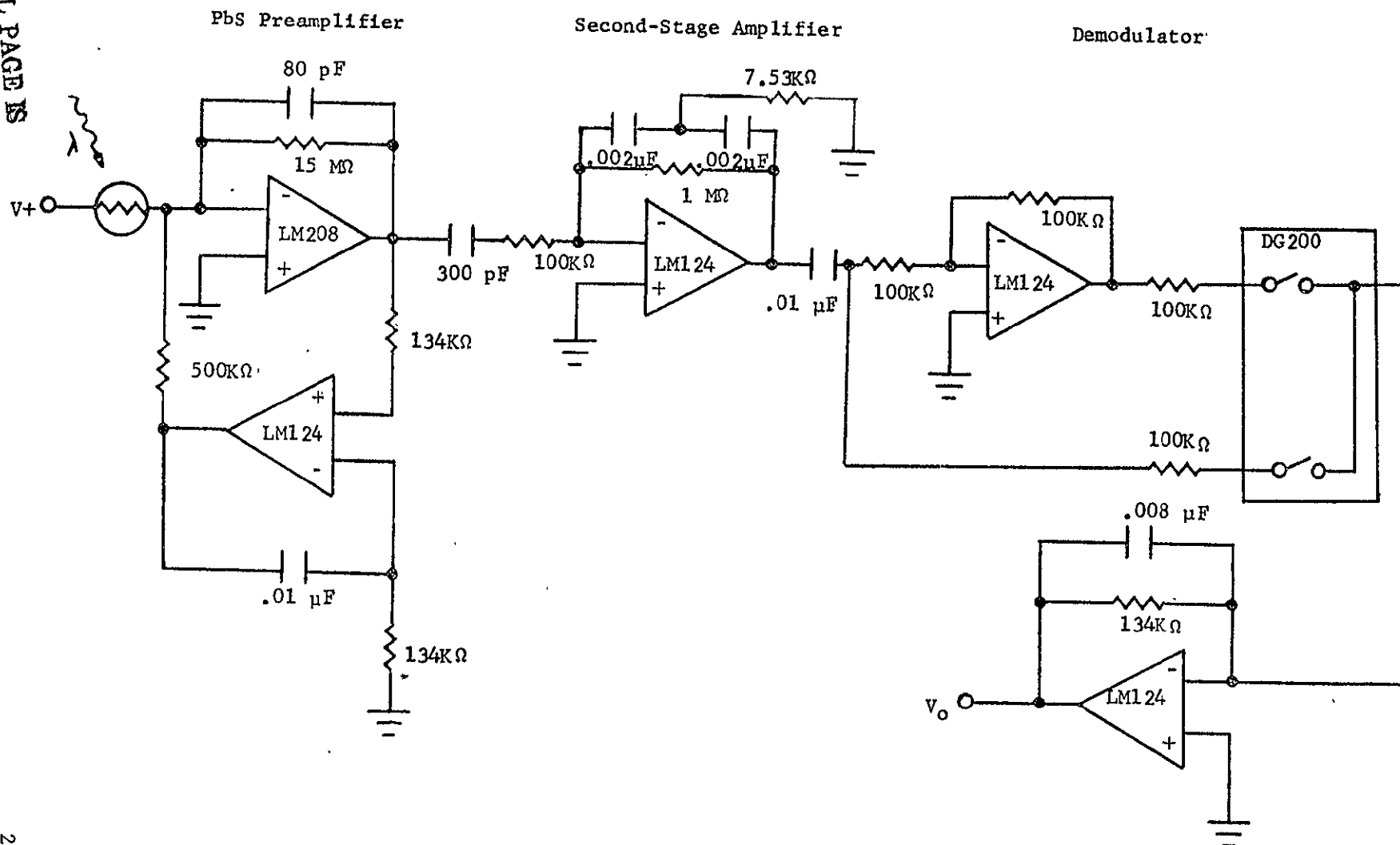


Figure 4. Lead Sulfide Detector Amplifier Circuitry for the AC-Chopped Light Design (Typical Channel)

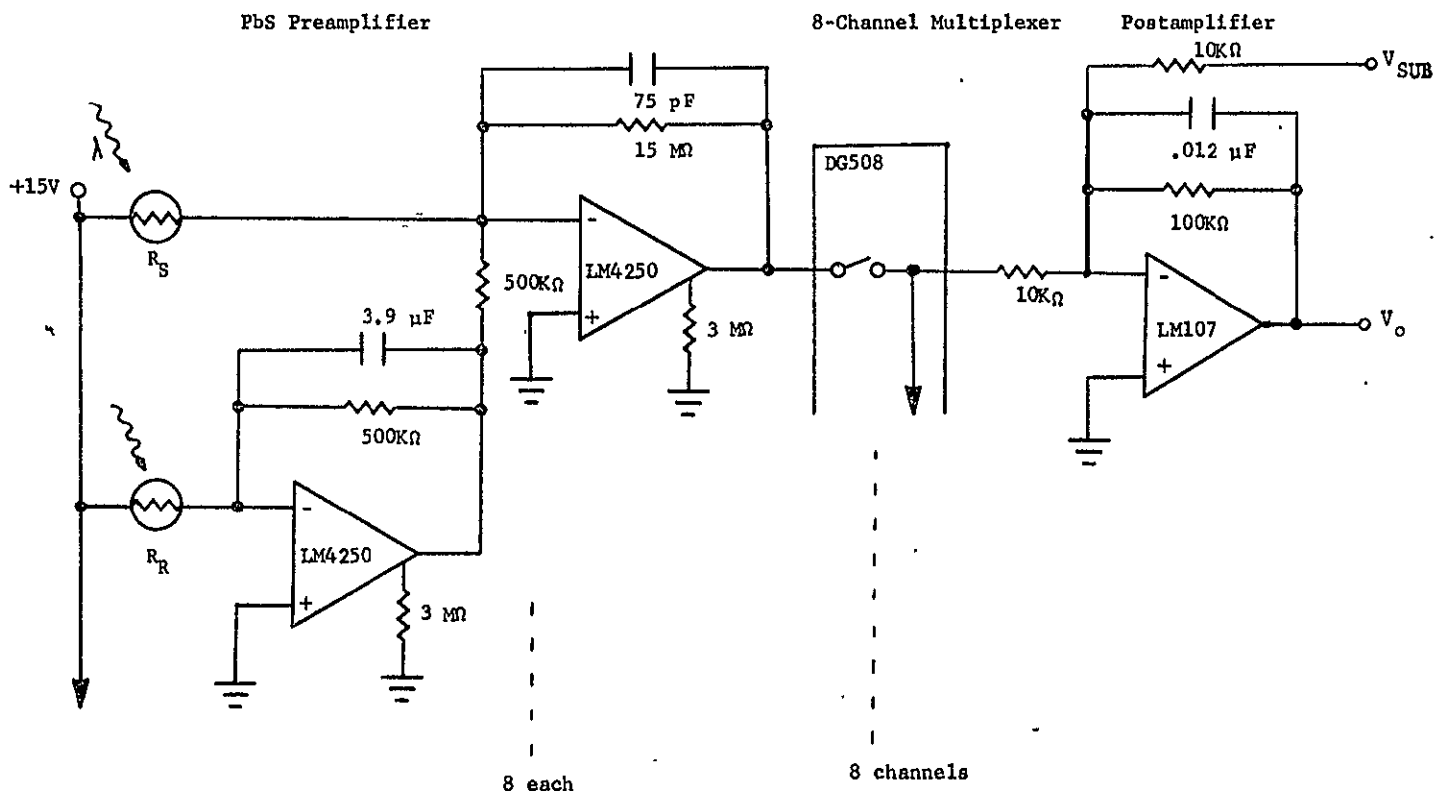


Figure 5. Lead Sulfide Detector Amplifier Circuitry for DC-Dual Detector Design

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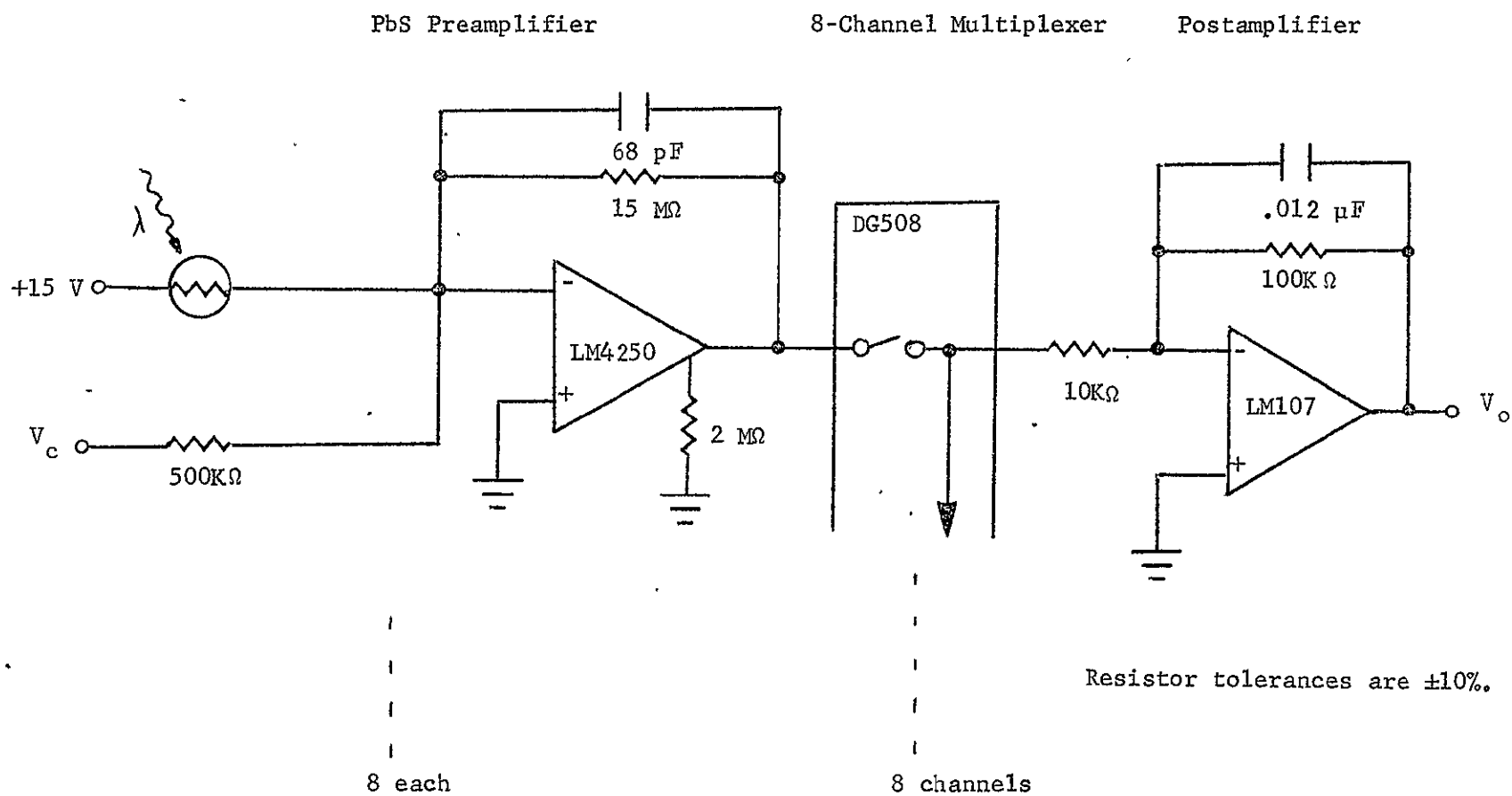


Figure 6. Lead Sulfide Detector Amplifier Circuitry for DC-Single Detector Design

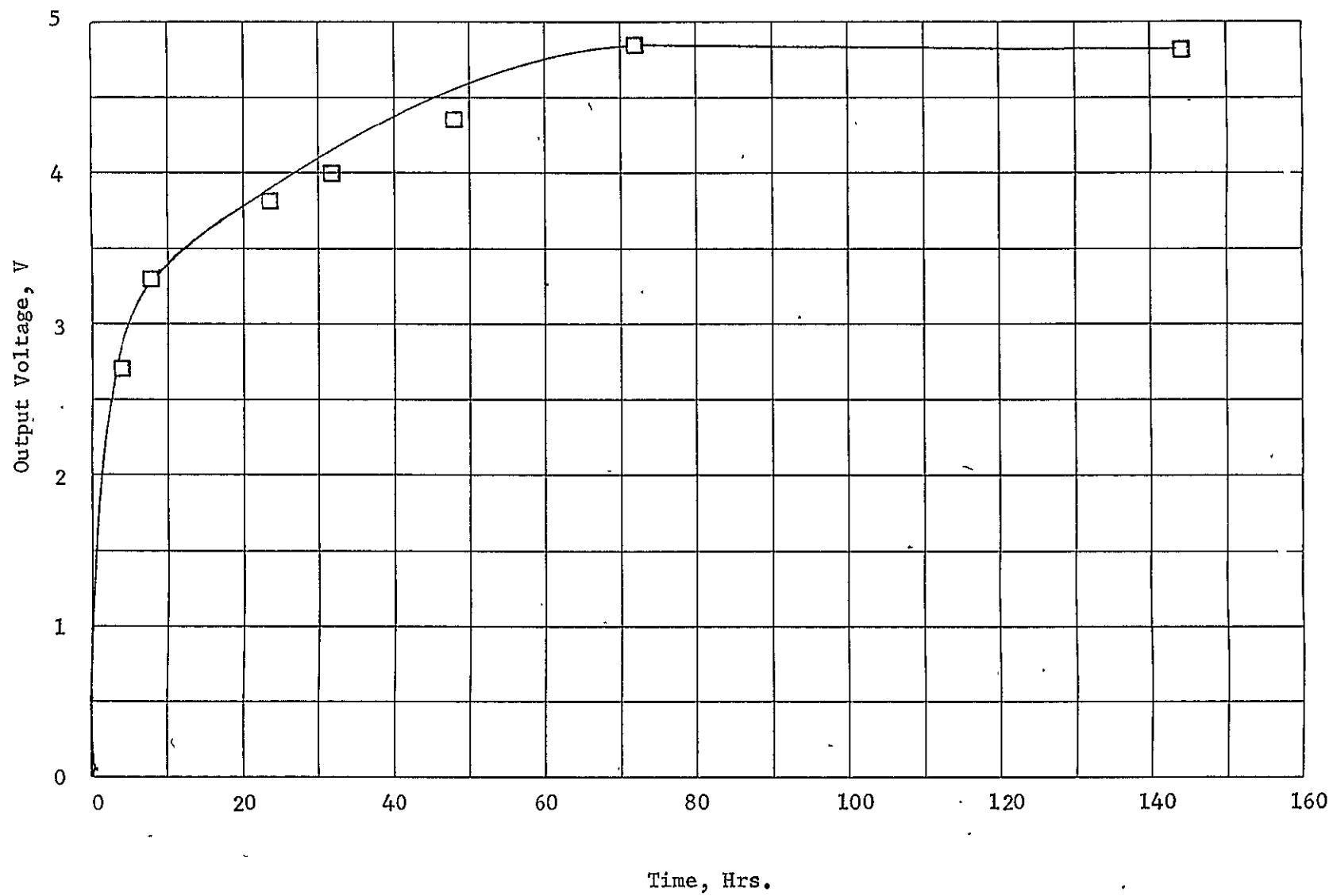


Figure 7. Room Ambient DC Drift of Dual Detector Breadboard

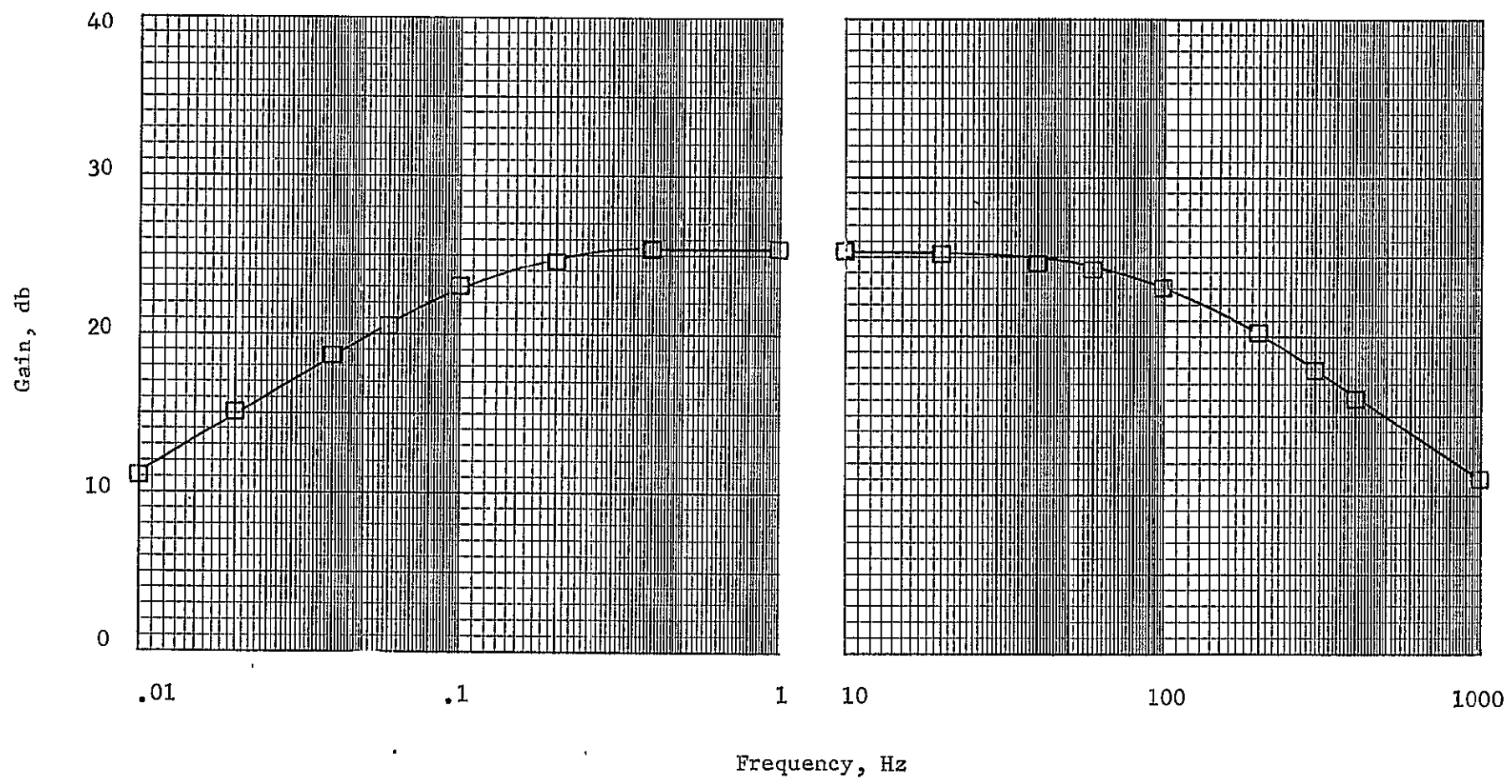


Figure 8. Frequency Response of PSA Dual Detector Breadboard

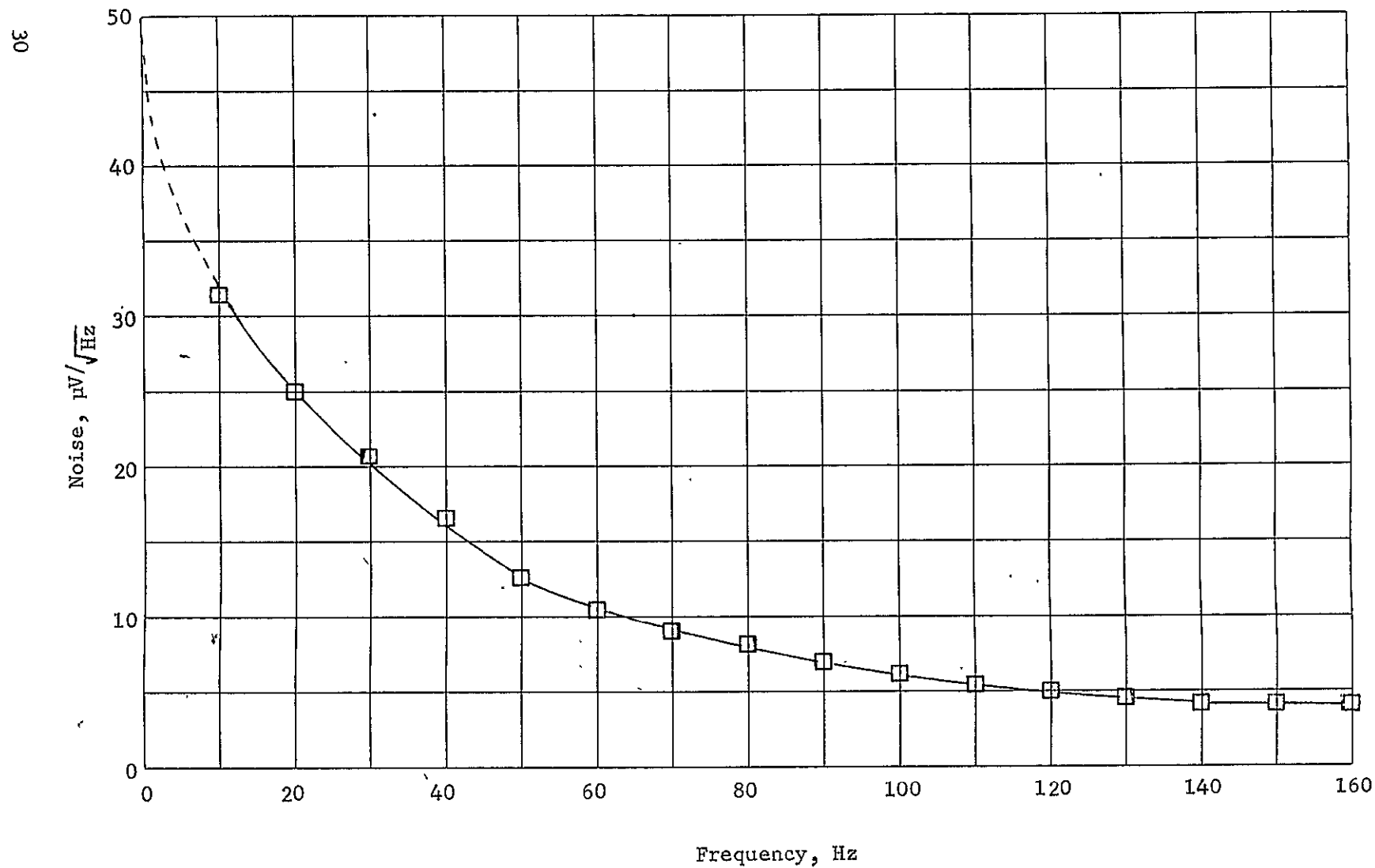
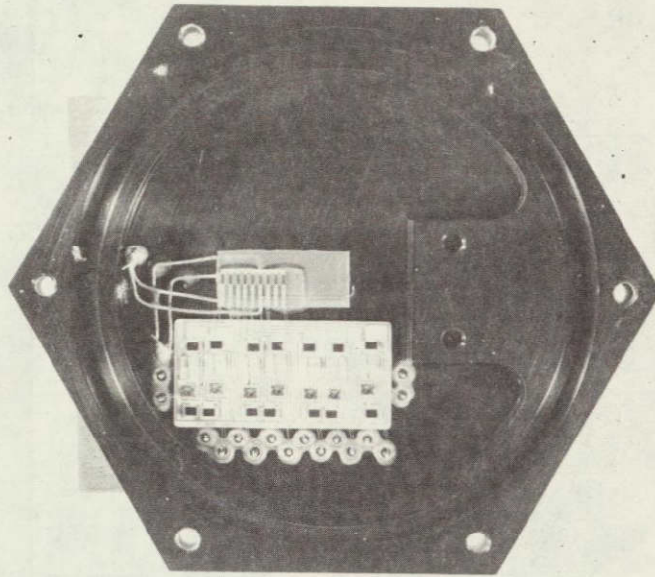
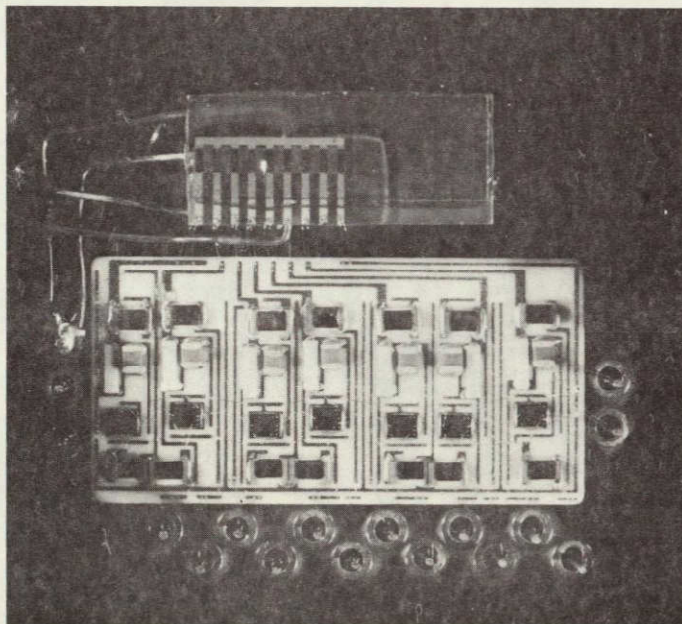


Figure 9. Noise Characteristics of Dual Detector Breadboard



(a) Prototype PSA



(b) Electronics and PbS Detector Array

Figure 10. Hybrid Prototype Photo Sensor Array

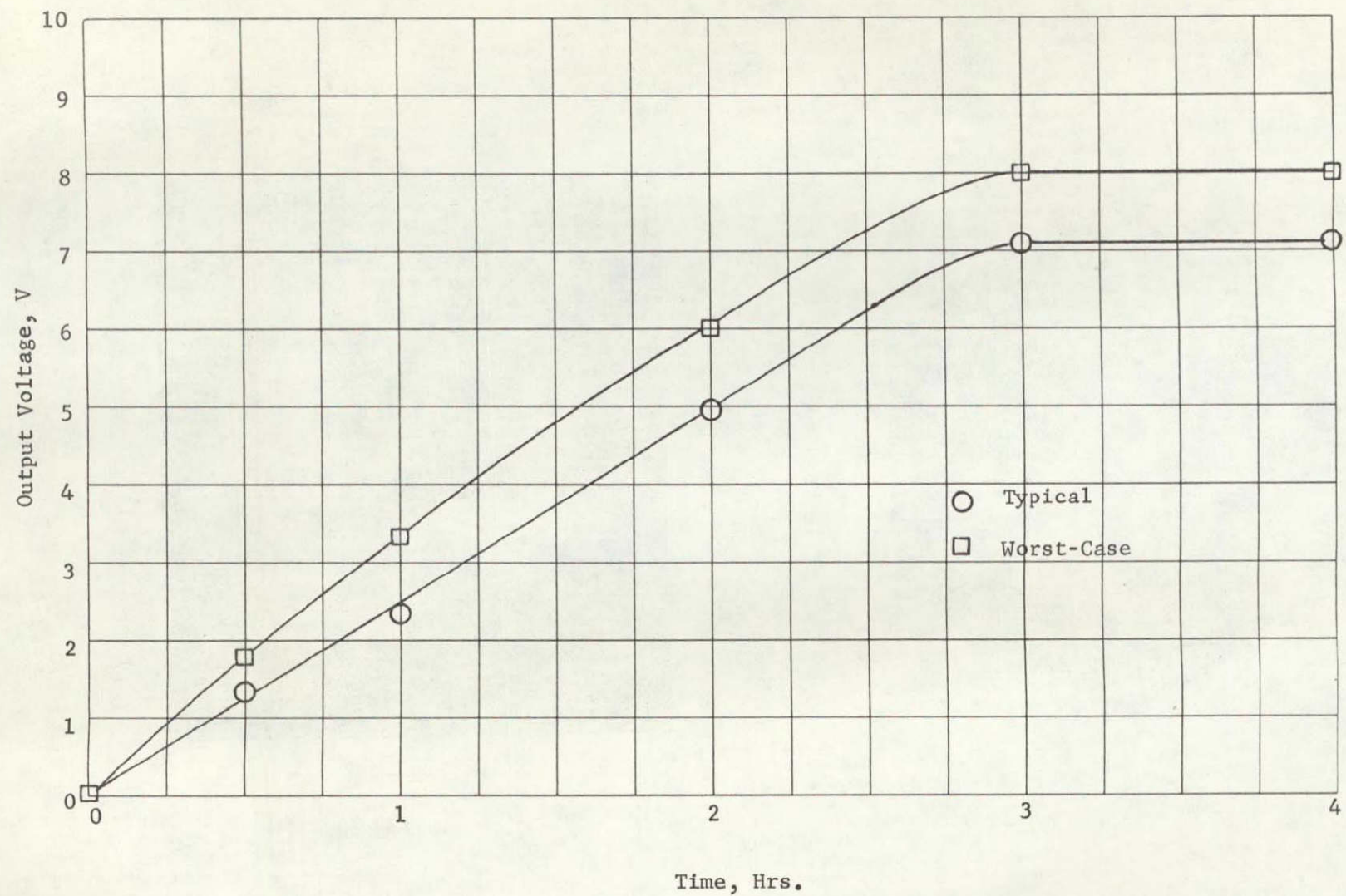


Figure 11. Controlled Ambient (+20 °C) DC Drift of PSA Single Detector Hybrid Microcircuit

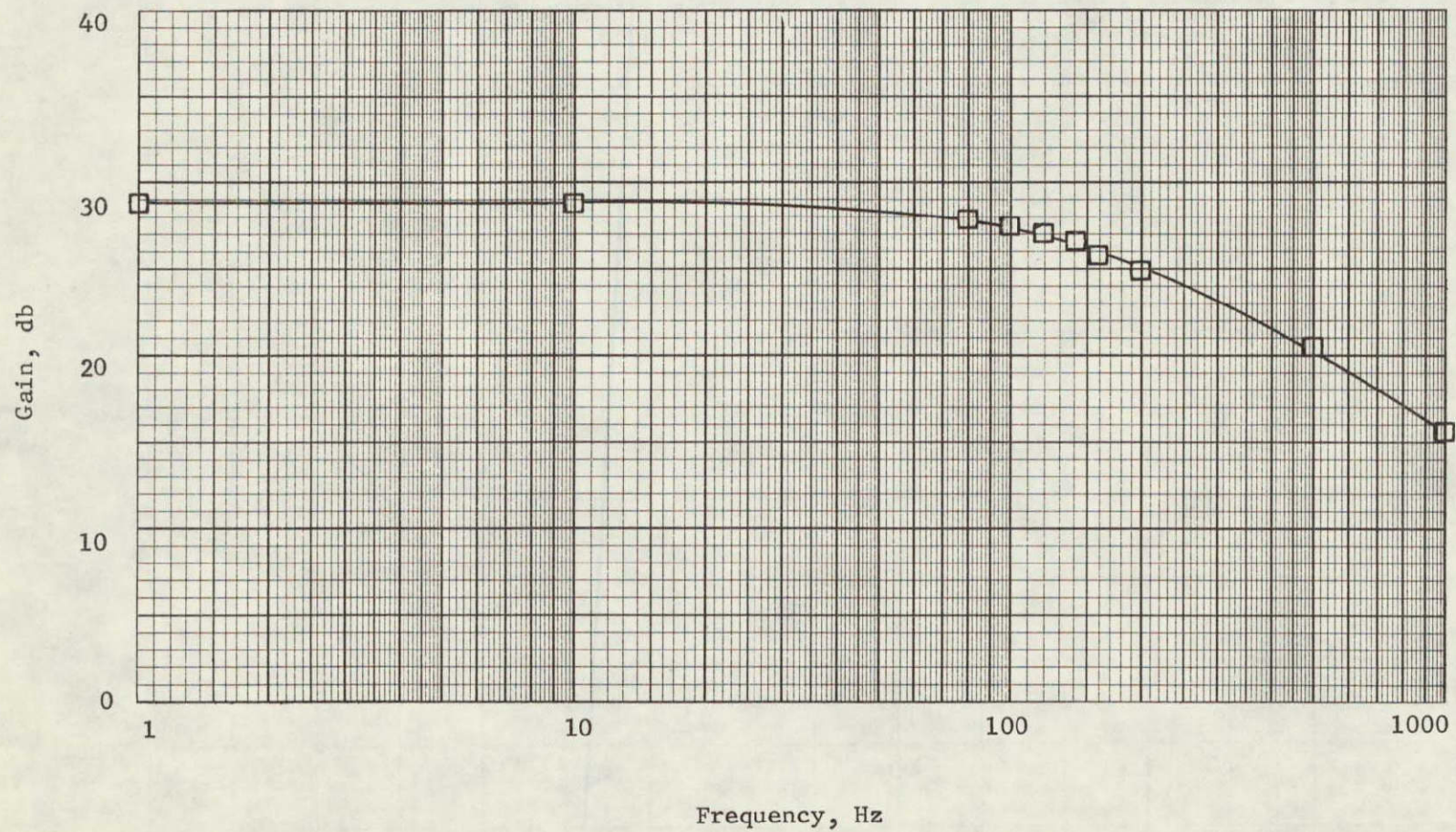


Figure 12. Frequency Response of PSA Single Detector Hybrid Amplifier

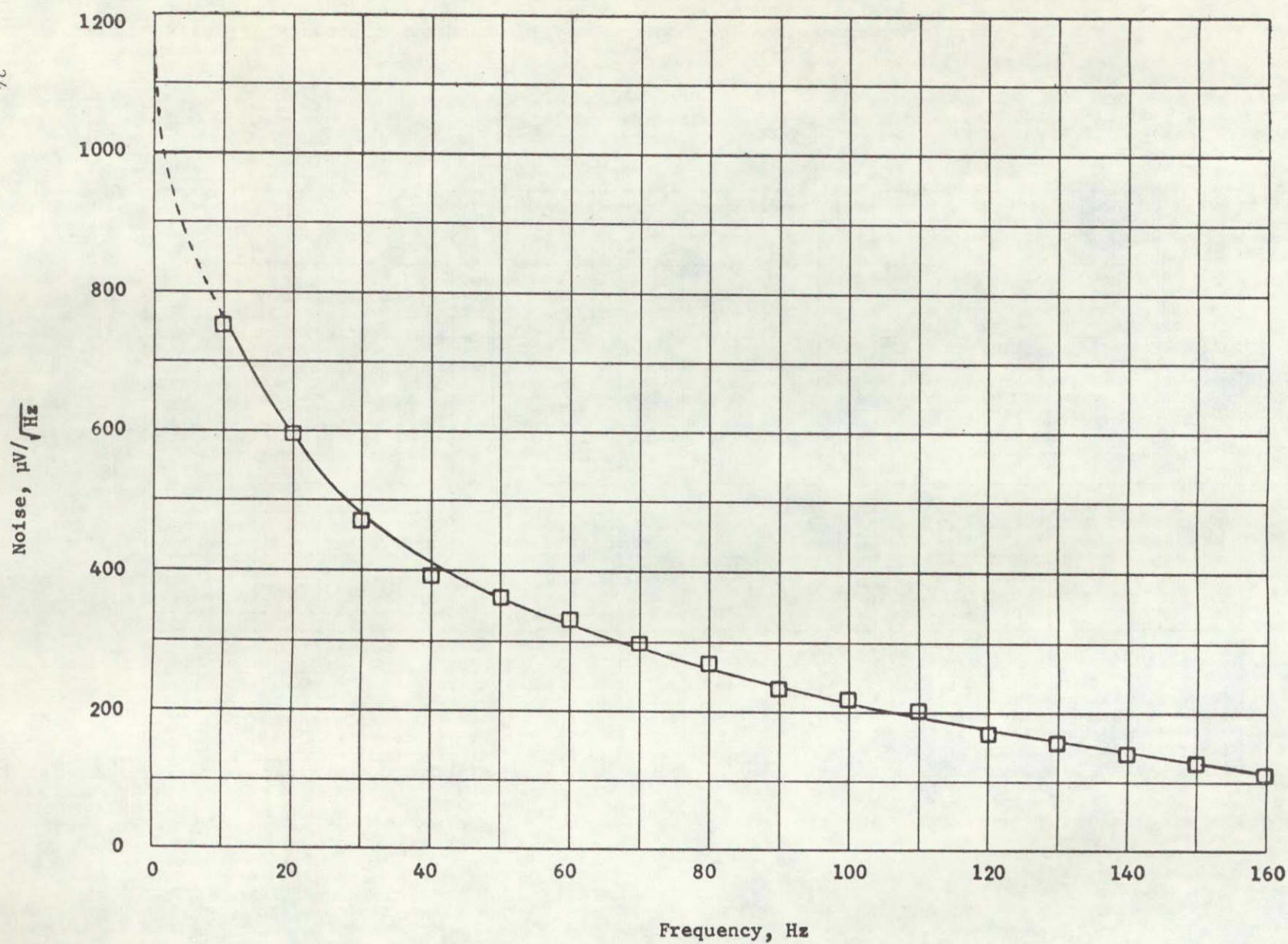
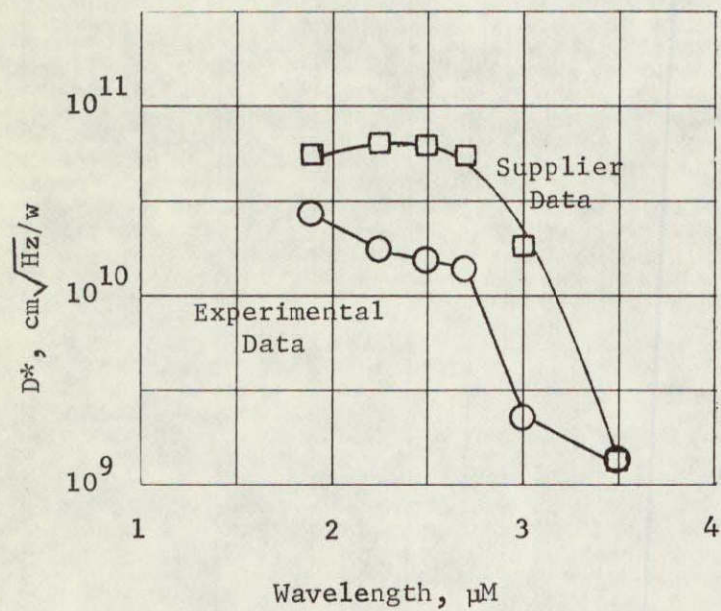
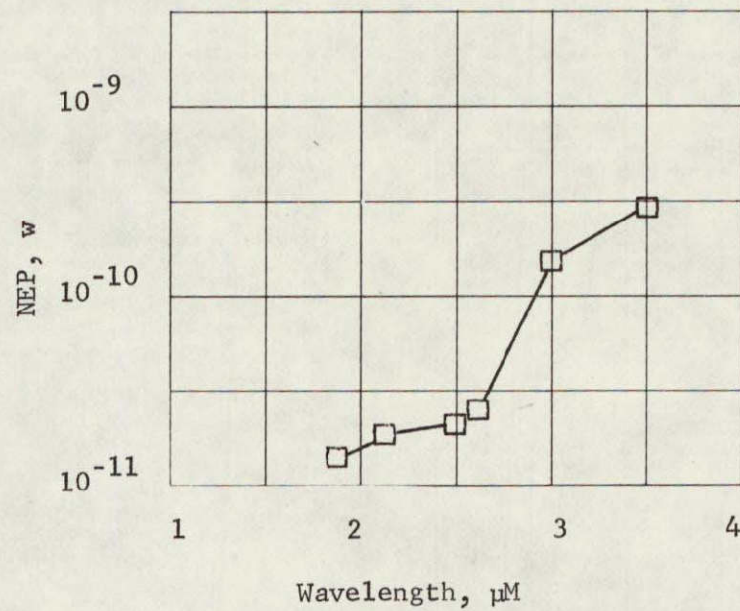


Figure 13. Noise Characteristics of PSA Single Detector Hybrid



a) Detectivity



b) Noise equivalent power

Figure 14. Spectral response of PSA single detector (PbS) hybrid

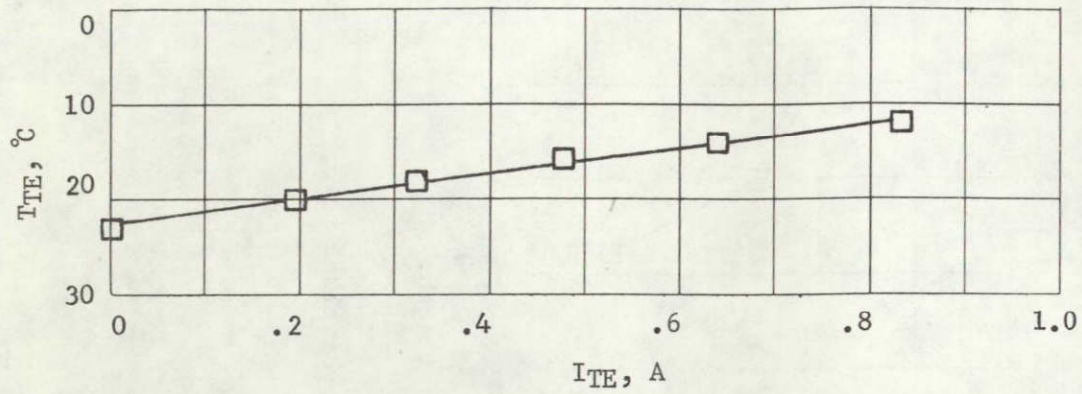


Figure 15. Thermoelectric Cooler Thermal Characteristics

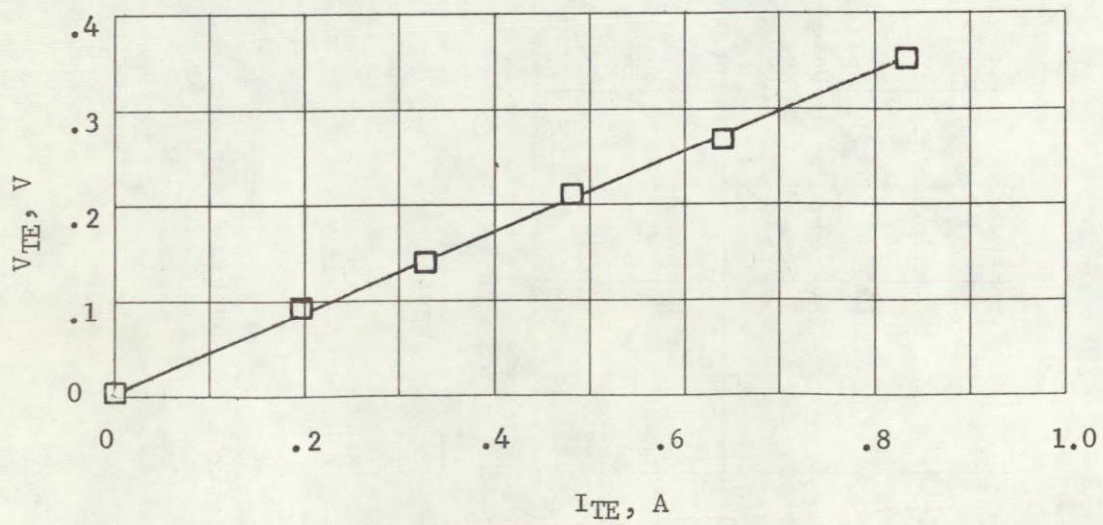


Figure 16. Thermoelectric Cooler Electrical Characteristics

APPENDIX A

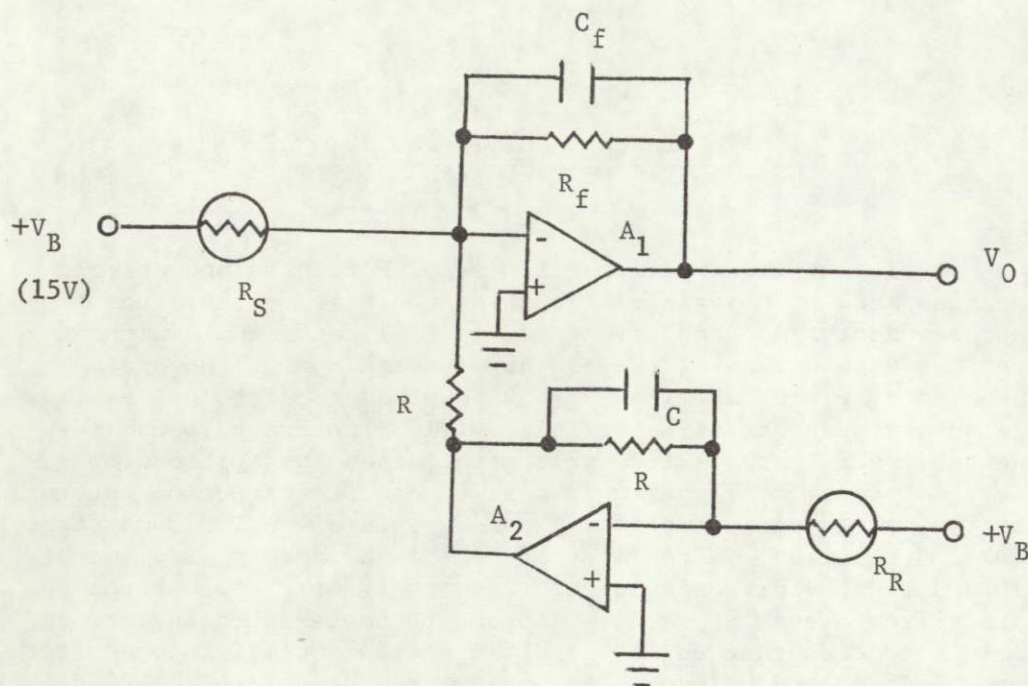
TOLERANCE ANALYSIS OF THE DC-DUAL DETECTOR PBS PREAMPLIFIER

A circuit for amplifying low level PbS photoconductive changes must be capable of rejecting the bias current from the photoconductor and amplifying low level light signals without serious degradation of the signal-to-noise ratio. Several schemes have been examined for DC coupled circuits (or very low frequency response circuits) that will be tolerant to the inherent instability of lead salt detectors. The application does not specifically allow a dark sample DC correction internal to the hybrid circuit or the problem would suggest a more straightforward solution. With the self-contained idea in mind, we are forced to consider using a reference cell (dark) to balance the drift from a sensing cell. This can be implemented in various bridge type approaches with a differential amplifier to monitor both legs of the bridge. One would like to achieve the DC drift cancellation of the bridge (and the power supply rejection) while avoiding increased I/F noise from the reference cell. After studying this problem, we found that this requires a considerably more complex amplifier than space would permit in an 8-channel hybrid circuit. The following circuit was selected because it is similar to the circuit that Bill Patterson suggested for a chopped light version of the PSA and will allow almost the same artwork to be used for both the DC and AC PSA. It does not provide any advantage for power supply rejection but does eliminate the noise contribution from the reference cell. From an amplifier noise contribution consideration, it is similar to an instrumentation amplifier and a standard bridge configuration circuit.

PbS Preamplifier

This circuit uses the reference cell and amplifier to provide an equal and opposite current into the summing node of A_1 . Current flowing the feedback resistor is only signal current. The capacitor, C, can be made relatively large to reject most of the low frequency noise of the reference cell. As will be shown, without feedback, the long term drift between the reference and sensing cells is still critical to maintain dynamic range for the amplifier output and the sample and subtract circuitry that follows the preamplifier. Preamplifier circuitry is shown below.

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Bias Current Rejection

The gain to the bias supply is:

$$V_O = \left(\frac{-V_B}{R_S} + \frac{V_B}{R_R} \right) R_f$$

Therefore, if the two cells are equal in resistance, the output is zero. The sensitivity to either cell drifting and also the low frequency gain is:

$$V_O = V_B \left(\frac{-R_f}{R_S + \Delta R_S} + \frac{R_f}{R_R} \right) = V_B \left[\frac{-R_f R_R + R_f (R_S + \Delta R_S)}{R_R (R_S + \Delta R_S)} \right]$$

with $V_B = 15$ volts, $R_S = R_R$ nominally:

$$V_O = 15 \frac{R_f \Delta R_S}{R_R (R_S + \Delta R_S)}$$

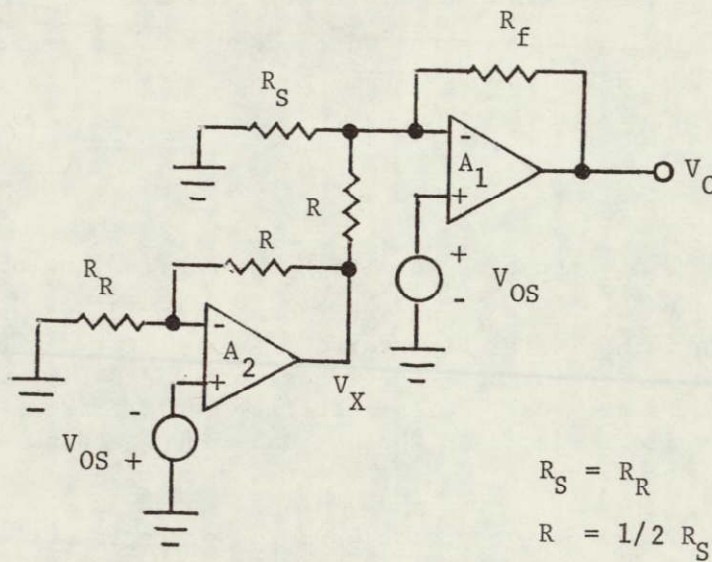
If we set the preamp gain for $R_f = 15 R_S$ which is probably on the low side from a signal-to-noise ratio consideration, but may be all that is practical from a DC drift standpoint, the output signal is:

$$V_0 = 225 \frac{\Delta R_S}{R_S + \Delta R_S}$$

A 1% relative drift produces 2.25 volts at the output. The system could probably stand 3 to 5% drift and still allow dark signal subtraction in the next stage. One of the primary tasks then, will be to empirically determine the sensor relative drift over time and environment. The final design will be a compromise between preamp gain and DC drift, as we expected.

Electronic Drift

Worst-case offset voltage occurs with maximum and opposite polarity offset voltages:



$$V_0 = V_{OS} \left(1 + \frac{R_f}{R_S} + \frac{R_f}{R} \right) - V_X \frac{R_f}{R}$$

$$V_X = - V_{OS} \left(1 + \frac{R}{R_R} \right)$$

$$V_0 = V_{OS} \left(1 + 2 \frac{R_f}{R_S} + 2 \frac{R_f}{R} \right)$$

For an LM4250 op amp over its full temperature range, the offset voltage is 5 mV maximum. With R set equal to $\frac{1}{2} R_S$, this contributes the following output voltage:

$$V_O = 5 \text{ mV} \left[1 + 2 (15) + 2 (30) \right] = 455 \text{ mV}$$

The bias current for the LM4250 flows out of the input terminal since it uses a PNP input stage. Bias current in the pre-amplifier is somewhat self-compensating for this amplifier configuration. The output voltage due to bias current is maximum when the bias current of A_2 is zero and the bias current of A_1 is maximum.

$$V_O = (-I_B) R_f$$

I_B is 15 nanoamps at room temperature (typical with $R_{SET} = 3 \text{ M}\Omega$). DC beta typically varies by about three over 100°C range, therefore, the maximum output change is:

$$\begin{aligned} V_O &= -(45 \times 10^{-9}) (15 \times 10^6) \\ &= 0.68 \text{ volts} \end{aligned}$$

This is an extreme worst-case condition since A_2 cannot operate without bias current but a significant change could appear at the output due to amplifier bias current. Amplifier drifts may be minimized by screening but sensor resistance and circuit resistance drift is critical. High stability resistors are required for R while it is found that R_f can change and not produce a dark level change.

The bias supply gain equation is more correctly written as:

$$V_O = \left(-\frac{V_B}{R_S} + V_B \frac{R}{R_R} \times \frac{1}{R} \right) R_f$$

The output voltage change with respect to R is found to be:

$$\left. \frac{V_O}{\Delta R} \right| = 225 \frac{\Delta R}{R}$$

which is the same sensitivity that the sensors produce. Thick film resistors of moderate geometry have been found to be more

stable than with minimum geometry. Typically they are stable within 1% over life. Printed simultaneously, the drift would be in the same direction with minimum consequences. Thin film produces more stable resistors--less than .2% drift over life with excellent tracking. The use of thin film would, however, require considerable "real estate" usage to achieve a value of several hundred thousand ohms (required because of Johnson noise considerations). The thin film process produces stable 200 ohms per square sheet resistance at +20 ppm/°C. Thick film is more typically ± 100 ppm/°C and would employ, for this case, 100K Ω per square ink. Thick film sheet resistances much above this suffer from static discharge damage and voltage coefficient problems. For a 100°C change, there could be a 1% change in R resulting in a 2.25 volt change in preamplifier output voltage if tracking did not occur. These resistors will be printed simultaneously with identical geometry to minimize relative drift. For analysis purposes we shall assume $\Delta R \leq .1\%$, therefore $\Delta V_{OUT} \leq .225$ volts.

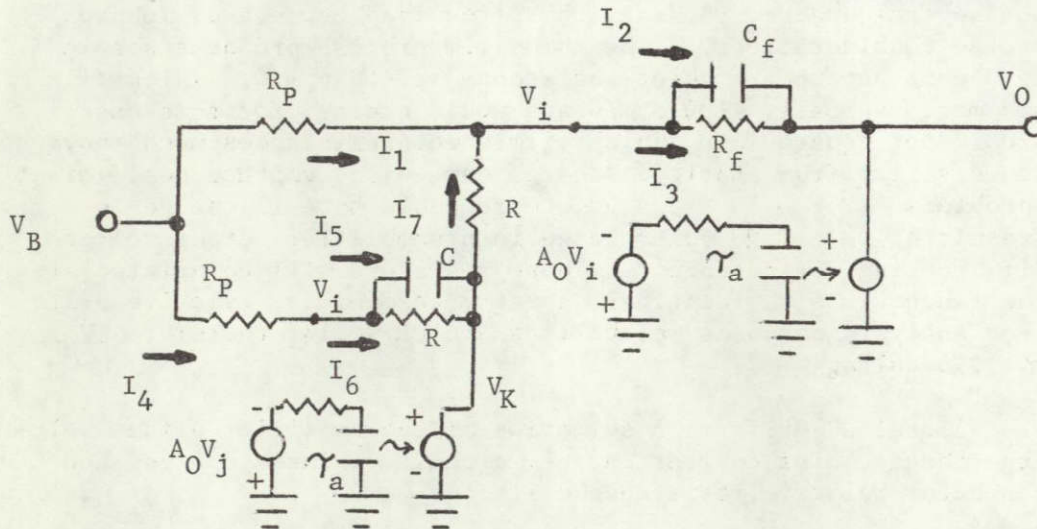
Total DC drift is a summation of the amplifier offset voltage change, bias current change, circuit resistor changes and detector relative resistance drift.

$$\begin{aligned}\Delta V_{OUT} &= \Delta V_{OUT_OFFSET} + \Delta V_{OUT_BIAS} + \Delta V_{OUT_CIRCUIT\ R} \\ &\quad + \Delta V_{OUT_DETECTOR} \\ &\cong 455\text{ mV} + 680\text{ mV} + 225\text{ mV} + \Delta V_{OUT_DETECTOR} \\ &\cong 1.36\text{ Volt} + \Delta V_{OUT_DETECTOR}\end{aligned}$$

This output change will consume most of the dynamic range capability of the post amplifier unless the post amplifier gain is low. We desire to achieve a gain of 10. Obviously, the gain must be around 2 to 3 if we find 1.36 volts of preamplifier output drift to be added to the detector drift. An alternative is to allow post amplifier DC correction with a sample and hold subtraction controlled by the camera shutter. This capability will be included in the hybrid. In conclusion, even if we find negligible drift error except for detector relative drift, the design will probably require less than 1% relative detector drift and a post amplifier stage gain of 3.

AC Analysis

A transfer function exists for power supply noise and ripple when we use a capacitor in the reference circuit. The power supply will require good regulation and heavy decoupling. The model for the AC power supply noise gain is:



This model neglects output resistance of the amplifiers and amplifier second pole break frequencies (a computer analysis will provide this).

$$I_1 = \frac{V_B - V_i}{R_P} = I_2 + I_3 - I_7$$

$$I_7 = \frac{V_K - V_i}{R}$$

$$I_4 = \frac{V_B - V_i}{R_P} = I_5 + I_6$$

$$I_4 = \frac{V_B - V_i}{R_P} = (V_j - V_K) \left(sC + \frac{1}{R} \right) = (V_j - V_K) \left(\frac{1 + s\tau}{R} \right)$$

Therefore,

$$V_K = - \left[\frac{R}{1 + s\tau} \frac{V_B - V_i}{R_P} - V_j \right]$$

$$= - \left[\frac{R}{R_P} \times \frac{V_B - V_j}{1 + s\tau} - V_j \right]$$

where $\tau = RC$

$$V_j = - V_K \frac{1 + s\tau_a}{A_0}$$

where $\tau_a = \text{Ampl. first pole}$

and $A_0 = \text{open loop gain}$

Therefore,

$$V_K = - \frac{R}{R_P} \left[\frac{V_B}{1 + s\tau} + \frac{V_K (1 + s\tau_a)}{(1 + s\tau) A_0} - V_K \frac{(1 + s\tau_a)}{A_0} \right]$$

and, after reduction:

$$V_K = - \frac{R V_B}{R_P (1 + s\tau) + \frac{R}{A_0} (1 + s\tau_a) - \frac{R}{A_0} (1 + s\tau_a) (1 + s\tau)}$$

with reasonable open loop gain, A_0 :

$$V_K (s) = - V_B \frac{R}{R_P} \frac{1}{1 + s\tau}$$

Making the same assumption, that A_0 is a large value:

$$V (s) = - (I_2 + I_3) \frac{R_f}{1 + s\tau_f} = - (I_1 + I_7) \frac{R_f}{1 + s\tau_f}$$

$$= - \left(\frac{V_B}{R_P} + \frac{V_K}{R} \right) \frac{R_f}{1 + s\tau_f}$$

$$= - \left[\frac{V_B}{R_P} - \frac{V_B}{R} \times \frac{R}{R_P} \times \frac{1}{1 + s\tau} \right] \frac{R_f}{1 + s\tau_f}$$

$$A (s) = \frac{V_O (s)}{V_B (s)} = - \frac{R_f}{R_P} \times \frac{s\tau}{(1 + s\tau_f) (1 + s\tau)}$$

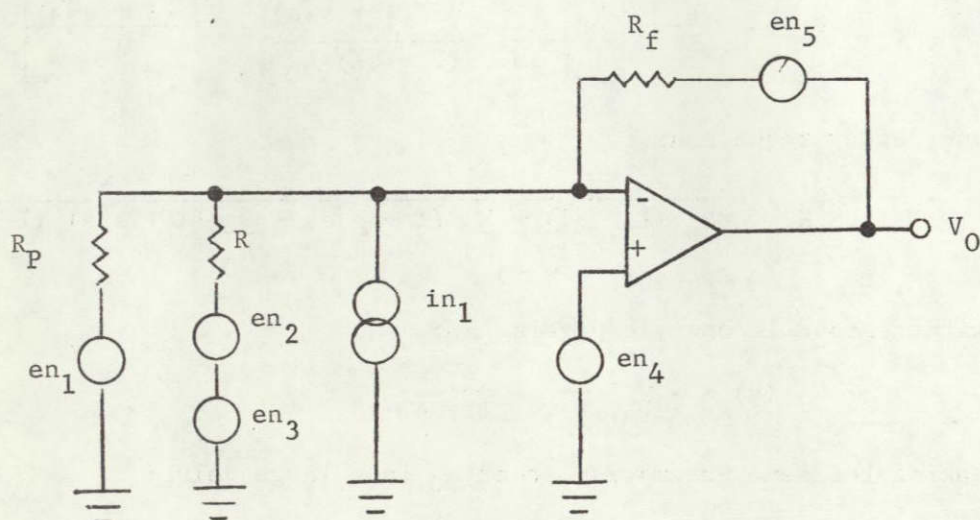
where $\tau_f = R_f C_f$

Thus, the gain to power supply noise has a bandpass filter form. Since the sample and subtract system has a low frequency cutoff of approximately $1/2 T_S$, where T_S is the sample period, we could set the amplifier time constant,

$$= T_S \cong 12 \text{ seconds}$$

similar to the Viking camera shutter speed. The upper break frequency is the bandwidth of 140 Hz.

Noise Model



en_1 = Detector noise

en_2 = Johnson noise from R

in_1 = Shot noise from bias current of amplifier

en_4 = Amplifier noise

en_5 = Johnson noise from R_f

en_3 = Reference amplifier noise

Noise Analysis

Because the circuit is operating at low frequencies, we are subject to $1/f$ noise in both amplifiers and in the detector. The noise source, en_3 , is the noise from the reference amplifier. It has both $1/f$ and white noise components but only has a gain of unity for frequencies above the sample and subtract low frequency limitation. The output RMS voltage noise density is:

$$V_{No}^2 = R_f \left[in_1^2 + \left(\frac{en_1}{R_P} \right)^2 + \left(\frac{en_2}{R} \right)^2 + \left(\frac{en_3}{R} \right)^2 \right] + (en_4)^2$$

$$\left(1 + R_f \times \frac{R + R_P}{R \times R_P} \right)^2 + 4KT R_f$$

The dominant terms will be the amplifier and detector noise since both are predominantly characterized by I/F noise. Bias current noise is:

$$in_1^2 = 2q I_{bias} = 2 \times 1.6 \times 10^{-19} \times 15 \times 10^{-9} =$$

$$4.8 \times 10^{-27} \left(\frac{\text{amps}}{\sqrt{\text{Hz}}} \right)^2$$

Johnson noise current for resistor R is:

$$\left(\frac{en_2}{R} \right)^2 = \frac{4KT}{R} = \frac{4 \times 1.38 \times 10^{-23} \times 300}{5 \times 10^5} =$$

$$3.31 \times 10^{-26} \left(\frac{\text{amps}}{\sqrt{\text{Hz}}} \right)^2$$

Reference amplifier noise is:

$$\left(\frac{en_3}{R} \right)^2 = \frac{50 \times 10^{-9}}{5 \times 10^5}^2 = 1 \times 10^{-26} \left(\frac{\text{amps}}{\sqrt{\text{Hz}}} \right)^2$$

The detector noise contribution is:

$$\left(\frac{en_1}{R_P} \right)^2 = \frac{K_N \tau V_{APP}}{D^*}^2$$

for Optoelectronics Detectors

$$D^* = 10^{11} \text{ (1 KHz)}, 8 \times 10^9 \text{ (1 Hz)}$$

$$\tau = 350 \text{ } \mu\text{s}$$

$$K_N = 14 \frac{\text{cm}}{\text{W-s-}\Omega}$$

$$V_{APP} = 15\text{V}$$

$$R_P = 10^6 \Omega$$

$$\left(\frac{en_1}{R_P}\right)^2 = \left(\frac{14 \times 350 \text{ } \mu\text{s} \times 15\text{V}}{8 \times 10^9}\right)^2 = 8.45 \times 10^{-22} \left(\frac{\text{amps}}{\sqrt{\text{Hz}}}\right)^2$$

The signal amplifier noise is multiplied by gain as follows:

$$en_4^2 \left[1 + R_f \times \frac{R + R_P}{R \times R_P}\right]^2 = en_4^2 (1 + 15 \times 3)^2 = 2120 en_4^2$$

assuming $R_f = 15 \text{ M}\Omega$, $R = 500 \text{ K}\Omega$, and $R_P = 1 \text{ M}\Omega$. The amplifier noise voltage is $50 \text{ nV}/\sqrt{\text{Hz}}$ from 40 Hz and up. Below 40 Hz, it is characterized by I/F noise and reaches 400 nanovolts at 1 Hz. The noise density above 40 Hz is:

$$2120 en_4^2 = 5 \times 10^{-12} \left(\frac{\text{Volts}}{\sqrt{\text{Hz}}}\right)^2$$

The feedback resistor contributes Johnson noise at unity gain:

$$en_5^2 = 4KT R_f = 4 \times 1.38 \times 10^{-23} \times 300 \times 1.5 \times 10^7 = 2.5 \times 10^{-13} \left(\frac{\text{Volts}}{\sqrt{\text{Hz}}}\right)^2$$

Finding the output RMS noise amounts to performing the following integration on each noise gain function:

$$V_{OUT} \text{ (RMS)} = K \left[\frac{1}{2\pi} \int_0^\infty |f(jw)|^2 dw \right]^{1/2}$$

This can be approximated, however, by considering abrupt break frequencies. The amplifier bandwidth is assumed abrupt at 140 Hz. The low frequency cutoff is also assumed to be abrupt at .1 Hz. These first order approximations allow us to find the output RMS noise voltage as follows:

$$\begin{aligned} \overline{V}_{NO}^2 = & R_f^2 i_{n1}^2 (140 \text{ Hz}) + R_f^2 \left(\frac{en_2}{R} \right)^2 (140 \text{ Hz}) + \\ & 2120 (en_4)^2 (140 \text{ Hz} - 40 \text{ Hz}) + \\ & 2120 (en_4)^2 \frac{1}{1 \text{ Hz}} \left[\ln \frac{40}{.1} \right] + en_5^2 (140 \text{ Hz}) + \\ & R_f^2 \left(\frac{en_1}{R_P} \right)^2 \frac{1}{1 \text{ Hz}} \left[\ln \frac{140}{.1} \right] + R_f^2 \left(\frac{en_3}{R} \right)^2 (140 \text{ Hz}) \end{aligned}$$

The op amp noise, en_4 , at 1 Hz can be read from the data sheet for the LM4250 and is $400 \text{ NV}/\sqrt{\text{Hz}}$.

The detector noise at 1 Hz must be derived from the fact that D^* is varying versely proportional to frequency.

$$en_1^2 \propto \frac{K^2}{f} \propto \left[\frac{K_N \gamma V_{APP} R_P}{D^* (1 \text{ Hz})} \right]^2 \frac{1}{f}$$

D^* is about 8×10^9 at 1 Hz for a PbS detector of 350 μsec time constant with peak D^* equal to 10^{11} at 1000 Hz.

Below is a table of each noise contribution:

Source	Equation
Bias Current	$R_f^2 i_{n1}^2 \Delta f = (15 \text{ M}\Omega)^2 (4.8 \times 10^{-27}) (140) = (12 \text{ }\mu\text{V})^2$
Detector	$\frac{R_f^2}{R_P^2} en_1^2 (1 \text{ Hz}) \left[\ln \frac{140}{.1} \right] = (1.5 \times 10^7)^2 (8.4 \times 10^{-23}) (7.24) = (370 \text{ }\mu\text{V})^2$

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Source	Equation
Resistor, R	$\frac{R_f^2}{R^2} (en_2)^2 \Delta f = \frac{(1.5 \times 10^7)^2}{5 \times 10^5} (4) (1.38 \times 10^{-23})$ $(300) (140) = (32 \mu V)^2$
Resistor, R _f	$en_5^2 \Delta f = (4) (1.38 \times 10^{-23}) (300) (1.5 \times 10^7)$ $(140) = (5.9 \mu V)^2$
Amplifier Noise (40 Hz to 140 Hz)	$2120 en_4^2 (140-40) = (2120) (5 \times 10^{-8})^2 (100) =$ $(23 \mu V)^2$
Amplifier Noise (.1 Hz to 40 Hz)	$2120 en_{4,1 \text{ Hz}}^2 \left[\ln \frac{40}{.1} \right] = (2120) (4 \times 10^{-7})^2$ $(6) = (45 \mu V)^2$
Reference Amplifier Noise	$\frac{R_f^2}{R^2} (en_3)^2 (140) = (1.5 \times 10^7)^2 (10^{-26}) (140) =$ $(18 \mu V)^2$

$$\bar{V}_{NO} = \left[(12 \mu V)^2 + (370 \mu V)^2 + (32 \mu V)^2 + (5.9 \mu V)^2 + (23 \mu V)^2 + (45 \mu V)^2 \right]^{1/2} \cong 375 \mu V$$

From an inspection of the output RMS noise equation, it is apparent that detector noise alone dominates. This would indicate that a lesser grade amplifier could be used with little compromise to the signal-to-noise ratio.

Let us examine actual data sheet performance on detectors that have been purchased for breadboard testing. Two detectors, .025 x .025 were purchased from Optoelectronics, Inc. They were tested with 12 volts applied across the detector and a 1 MΩ bias resistor. The noise was 2.0 μV with a 10 Hz bandpass.

$$V_N = \frac{K_N \gamma R_P V_{APP}}{D^*} \sqrt{\ln \frac{600}{590}}$$

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The D^* is assumed to be 8×10^9 at 1 Hz

$$V_N = (9.2 \times 10^{-6}) (.1296) = 1.2 \mu V$$

or reasonably close to the tested value. This tends to confirm the fact that the detector noise derivation and D^* extrapolation is correct. The obvious choice for an amplifier is the LM4250 that was used in the Viking PSA. Specification limits show that power consumption is only 3m watt (for $I_{SET} = 10 \mu A$) and the noise is $50 \text{ NV}/\sqrt{\text{Hz}}$.

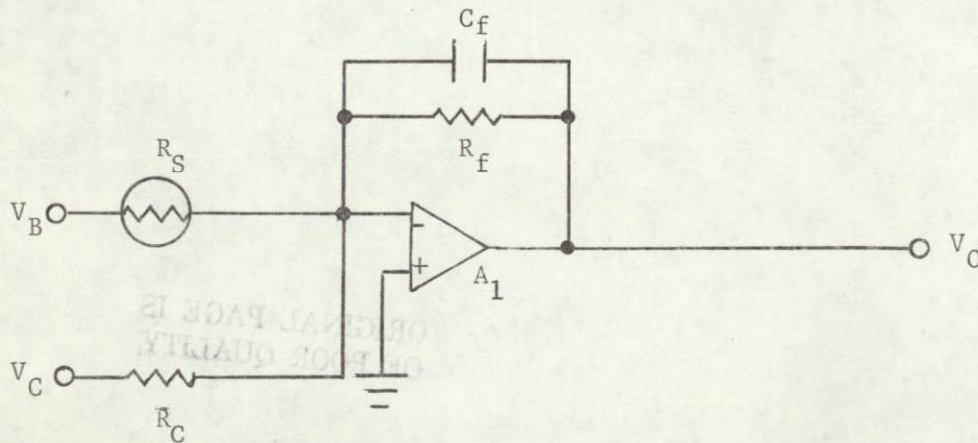
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APPENDIX B

TOLERANCE ANALYSIS OF THE DC-SINGLE DETECTOR PBS PREAMPLIFIER

A circuit for amplifying low level PbS photoconductive changes must be capable of rejecting the bias current from the photoconductor and amplifying low level light signals without serious degradation of the signal-to-noise ratio. Several schemes have been examined for DC coupled circuits (or very low frequency response circuits) that will be tolerant to the inherent instability of lead salt detectors. The following (single detector) circuit was selected because it eliminates the need for matched detector TCR's required by the dual detector approach previously investigated. It also will allow use of from eight to sixteen channels within the hybrid package. It will provide advantages in the areas of DC drift and error due to offset voltages. The primary disadvantage is that it requires dark signal correction by either dark signal sampling or application of a correction voltage, V_C .

Preamplifier Diagram



This circuit uses external correction voltage to provide an equal and opposite current into the summing node of A_1 . Current flowing in the feedback resistor is only signal current. The resistor, R_C , can be made equal to R_S at dark signal conditions

for optimal design while R_f and C_f are used to set amplifier gain and frequency response.

Bias Current Rejection

The gain to the bias supply is:

$$V_O = \left(-\frac{V_B}{R_S} - \frac{V_C}{R_C} \right) R_f$$

Therefore, if R_C and R_S are equal in resistance and $V_C = -V_B$, the output is zero. The sensitivity to cell drifting and also the low frequency gain is:

$$V_O = \frac{-R_f V_B}{R_S + \Delta R_S} - \frac{R_f V_C}{R_C} = \frac{-R_f R_C V_B - R_f (R_S + \Delta R_S) V_C}{R_C (R_S + \Delta R_S)}$$

with $V_B = -V_C = 15$ volts, $R_S = R_C$ nominally:

$$V_O = 15 \frac{R_f \Delta R_S}{R_C (R_S + \Delta R_S)}$$

If we set the preamp gain for $R_f = 30 R_S$ which may be more than is practical from a DC drift standpoint, the output signal is:

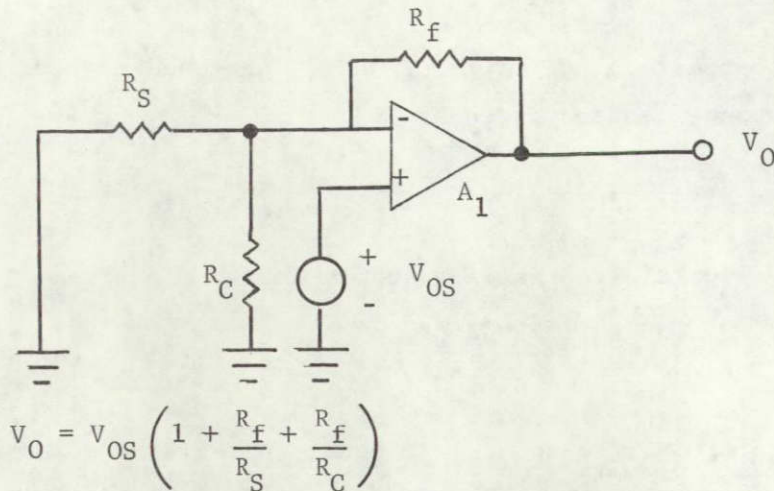
$$V_O = 450 \frac{\Delta R_S}{R_S + \Delta R_S}$$

A 1% relative drift produces 4.50 volts at the output. The system could probably stand 3 to 5% drift and still allow dark signal subtraction in a post amplifier stage. One of the primary tasks then will be to empirically determine the sensor relative drift over time and environment. The final design will be a compromise between preamp gain and DC drift, as we expected.

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Electronic Drift

Worst-case offset voltage occurs as follows:



For an LM4250 op amp over its full temperature range, the offset voltage is 5 mV maximum. With R_C set equal to R_S , this contributes the following output voltage:

$$V_O = 5 \text{ mV} (1 + 30 + 30) = 305 \text{ mV}$$

The bias current for the LM4250 flows out of the input terminal since it uses a PNP input stage. Bias current in the pre-amplifier is somewhat self-compensating for this amplifier configuration. The output voltage due to bias current is maximum when the bias current of A_1 is maximum.

$$V_O = (-I_B) R_f$$

I_B is 15 nanoamps at room temperature (typical with $R_{SET} = 2 \text{ M}\Omega$). DC beta typically varies by about three over 100°C range, therefore, the maximum output change is:

$$\begin{aligned} V_O &= -(45 \times 10^{-9}) (15 \times 10^6) \\ &= 0.68 \text{ volts} \end{aligned}$$

The output voltage change with respect to R_C is found to be:

$$\left. \frac{V_O}{\Delta R_C} \right|_{\Delta R_C} = 450 \frac{\Delta R_C}{R_C}$$

which is the same sensitivity that the R_s sensor produces. Thick film chip resistors of moderate geometry have been found to be more stable than with minimum geometry and will be used. Typically they are stable within .1% over life. For analysis purposes we shall assume $\Delta R \leq .1\%$, therefore, $\Delta V_{OUT} \leq .450$ volts.

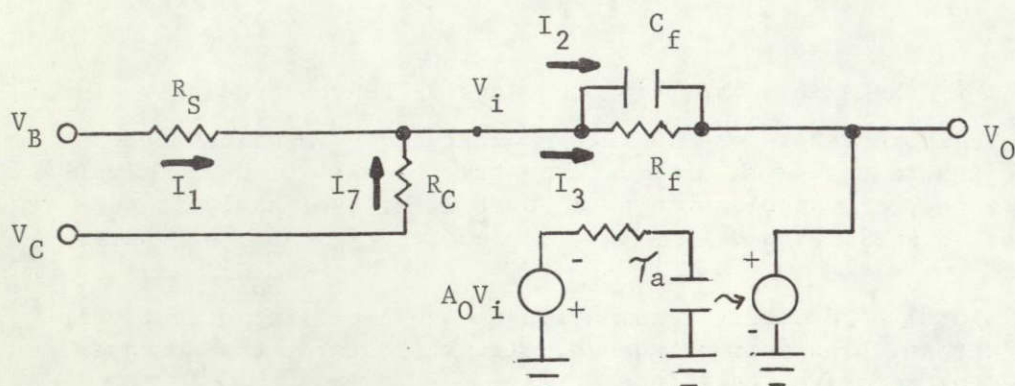
Total DC drift is a summation of the amplifier offset voltage change, bias current change, circuit resistor changes and detector relative resistance drift.

$$\begin{aligned}\Delta V_{OUT} &= \Delta V_{OUT_OFFSET} + \Delta V_{OUT_BIAS} + \Delta V_{OUT_CIRCUIT\ R} + \\ &\quad \Delta V_{OUT_DETECTOR} \\ &\cong 305\text{ MV} + 680\text{ MV} + 450\text{ MV} = \Delta V_{OUT_DETECTOR} \\ &\cong 1.44\text{ Volt} + \Delta V_{OUT_DETECTOR}\end{aligned}$$

It should be noted that since multiplexing and post amplification technology are well established, they will not be employed in the hybrid prototype test vehicle being produced. However, since they could be incorporated in future designs with minor circuit additions and package modifications, the following analyses may be of interest. The output calculated above will consume most of the dynamic range capability of a post amplifier unless the post amplifier gain is low. We desire to achieve a gain of 10. Obviously, the gain must be around 3 if we find 1.44 volts of preamplifier output drift to be added to the detector drift. An alternative is to allow post amplifier DC correction with a sample and hold subtraction controlled by the camera shutter. In conclusion, even if we find negligible drift error except for detector relative drift, the design will probably require less than 1% relative detector drift and a post amplifier stage gain of 3.

AC Analysis

A transfer function exists for power supply noise and ripple. The power supply will require good regulation and heavy decoupling. The model for the AC power supply noise gain is:



This model neglects output resistance of the amplifiers. Current equations for this circuit are as follows:

$$I_1 = \frac{V_B - V_i}{R_S} = I_2 + I_3 - I_7$$

$$I_7 = \frac{V_C - V_i}{R_C}$$

Therefore,

$$I_2 + I_3 = I_1 + I_7 = \frac{V_B - V_i}{R_S} + \frac{V_C - V_i}{R_C}$$

$$\begin{aligned} I_2 + I_3 &= (V_i - V_O) \left(sC_f + \frac{1}{R_f} \right) \\ &= (V_i - V_O) \frac{1 + s\gamma_f}{R_f} \\ &= V_O \frac{1 + s\gamma_f}{R_f} + V_i \frac{1 + s\gamma_f}{R_f} \end{aligned}$$

where $\gamma_f = R_f C_f$

$$\text{or } -V_O = \frac{(I_2 + I_3) - V_i \frac{1 + s\gamma_f}{R_f}}{\frac{1 + s\gamma_f}{R_f}}$$

and substituting for $V_i = -V_O \frac{1 + s\gamma_a}{A_0}$ and $I_2 + I_3$:

$$V_O = \frac{-V_O \frac{1 + s\tau_a}{A_O} \frac{1 + s\tau_f}{R_f} - \frac{V_B + V_O \frac{1 + s\tau_a}{A_O}}{R_S}}{\frac{1 + s\tau_f}{R_f} - \frac{V_C + V_O \frac{1 + s\tau_a}{A_O}}{\frac{R_C}{\frac{1 + s\tau_f}{R_f}}}}$$

with reasonably large open loop gain, A_O :

$$V_O(s) = -\left(\frac{V_B}{R_S} + \frac{V_C}{R_C}\right) \frac{R_f}{1 + s\tau_f}$$

and at $V_C = -V_B$:

$$V_O(s) = -V_B \frac{R_C - R_S}{R_S R_C} \frac{1}{1 + s\tau_f}$$

or,

$$A(s) = \frac{V_O(s)}{V_B(s)} = -\frac{R_C - R_S}{R_S R_C} R_f \frac{1}{1 + s\tau_f}$$

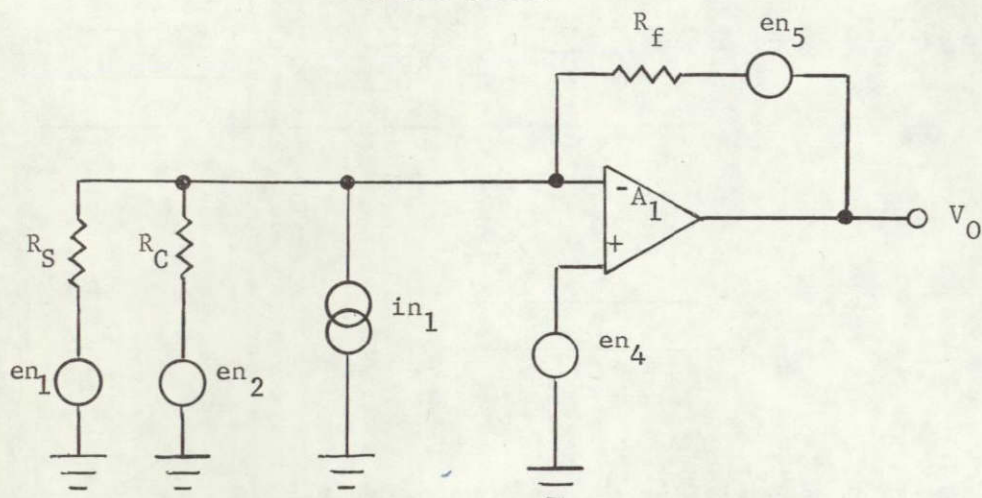
Thus, the gain to power supply noise has a typical amplifier form with an upper break frequency. Since the sample and subtract system has a low frequency cutoff of approximately $1/2 T_S$, where T_S is the sample period, we could set the amplifier time constant,

$$= T_S \cong 12 \text{ seconds}$$

similar to the Viking camera shutter speed. The upper break frequency is the bandwidth of 140 Hz.

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Noise Model



en_1 = Detector noise

en_2 = Johnson noise from R_C

in_1 = Shot noise from bias current of amplifier

en_4 = Amplifier noise

en_5 = Johnson noise from R_f

Noise Analysis

Because the circuit is operating at low frequencies, we are subject to I/F noise in both amplifier and detector.

The output RMS voltage noise density is:

$$V_{NO}^2 = R_f^2 \left[in_1^2 + \left(\frac{en_1}{R_S} \right)^2 + \left(\frac{en_2}{R_C} \right)^2 \right] + en_4^2 \left(1 + R_f \frac{R_C + R_S}{R_C R_S} \right)^2 + en_5^2$$

The dominant terms will be the amplifier and detector noise since both are predominantly characterized by I/F noise. Bias current noise is:

$$\begin{aligned} i_{n1}^2 &= 2q I_{\text{bias}} = 2 \times 1.6 \times 10^{-19} \times 15 \times 10^{-9} = \\ &= 4.8 \times 10^{-27} \left(\frac{\text{amps}}{\sqrt{\text{Hz}}} \right)^2 \end{aligned}$$

Johnson noise current for resistor R_C is:

$$\left(\frac{e_{n2}}{R_C} \right)^2 = \frac{4KT}{R_C} = \frac{4 \times 1.38 \times 10^{-23} \times 300}{5 \times 10^5} = 3.31 \times 10^{-26} \left(\frac{\text{amps}}{\sqrt{\text{Hz}}} \right)^2$$

The detector noise contribution is:

$$\left(\frac{e_{n1}}{R_S} \right)^2 = \left[\frac{K_N \gamma V_{\text{APP}}}{D^*} \right]^2$$

for Optoelectronics Detectors:

$$D^* = 10^{11} \text{ (1 KHz)}, 8 \times 10^9 \text{ (1 Hz)}$$

$$\gamma = 350 \text{ } \mu\text{S}$$

$$K_N = 14 \frac{\text{cm}}{2\text{-s-}\Omega}$$

$$V_{\text{APP}} = 15 \text{ V}$$

$$R_S = .5 \times 10^6 \Omega$$

The signal amplifier noise is multiplied by gain as follows:

$$e_{n4}^2 \left(1 + R_f \frac{R_C + R_S}{R_C R_S} \right)^2 = 3721 e_{n4}^2$$

assuming $R_f = 15 \text{ M}\Omega$, $R_C = 500 \text{ K}\Omega$ and $R_S = .5 \text{ M}\Omega$

The amplifier noise voltage is $50 \text{ NV}/\sqrt{\text{Hz}}$ from 40 Hz and up.

Below 40 Hz, it is characterized by I/F noise and reaches 400 nanovolts at 1 Hz.

The feedback resistor contributes Johnson noise at unity gain:

$$\begin{aligned} en_5^2 &= 4KT R_f = 4 \times 1.38 \times 10^{-23} \times 300 \times 1.5 \times 10^7 \\ &= 2.5 \times 10^{-13} \left(\frac{\text{volts}}{\sqrt{\text{Hz}}} \right)^2 \end{aligned}$$

Finding the output RMS noise amounts to performing the following integration on each noise gain function:

$$V_O \text{ (RMS)} = K \left[\frac{1}{2\pi} \int_0^\infty \left| f(jw) \right|^2 dw \right]^{1/2}$$

The output RMS noise voltage is:

$$\begin{aligned} \bar{V}_{NO}^2 &= R_f^2 in_1^2 (140 \text{ Hz}) + R_f^2 \left(\frac{en_2}{R_C} \right)^2 (140 \text{ Hz}) \\ &+ \int_0^{140} R_f^2 \left(\frac{en_1}{R_S} \right)^2 df + en_5^2 (140 \text{ Hz}) \\ &+ \int_0^{140} 2120 en_4^2 \left(\frac{1}{1 + 2\pi f \tau_f} \right)^2 df \end{aligned}$$

Two integrations must be performed to complete the noise calculations. The op amp noise contribution of en_4 is as follows:

$$\begin{aligned} \bar{V}_{NO}^2 &= \int_0^f 3721 (en_4)^2 \left(\frac{1}{1 + 2\pi f \tau_f} \right)^2 df \\ &= 3721 \left[(4 \times 10^{-7})^2 \int_0^{40} (1 + 2\pi f \tau_f)^{-2} df \right. \\ &\quad \left. + (5 \times 10^{-8})^2 \int_{40}^{140} (1 + 2\pi f \tau_f)^{-2} df \right] \\ &= 2.5 \times 10^{-8} \text{ V} = (157 \text{ } \mu\text{V})^2 \end{aligned}$$

where $\tau_f = 15 \text{ M}\Omega \times 68 \text{ pf}$

The detector noise must be derived from the fact that D^* is varying inversely proportional to frequency.

$$\left(\frac{en_1}{R_S}\right)^2 \propto \frac{K^2}{f} \propto \left[\frac{K_N \gamma V_{APP}}{D^*}\right]^2 \frac{1}{f}$$

D^* is about 8×10^9 at 1 Hz for a PbS detector of 350 μ sec time constant with peak D^* equal to 10^{11} at 1000 Hz.

$$\begin{aligned}\bar{V}_{NO}^2 &= \int_0^f R_f^2 \left(\frac{en_1}{R_S}\right)^2 df \\ &= 1.81 \times 10^{-7} \text{ V} = (426 \text{ } \mu\text{V})^2\end{aligned}$$

Below is a table of each noise contribution:

Source	Equation
Bias Current	$R_f^2 i_{n1}^2 \Delta f = (15 \times 10^6)^2 \times 4.8 \times 10^{-27} \times 140$ $= (12 \text{ } \mu\text{V})^2$
Detector	$\int_0^{140} R_f^2 \left(\frac{en_1}{R_S}\right)^2 df = (426 \text{ } \mu\text{V})^2$
Resistor, R_C	$\frac{R_f^2}{R_C^2} en_2^2 \Delta f = \frac{(1.5 \times 10^7)^2}{5 \times 10^5} (4) (1.38 \times 10^{-23})$ $(300) (140) = (32 \text{ } \mu\text{V})^2$
Resistor, R_f	$en_5^2 \Delta f = 4 (1.38 \times 10^{-23}) (300) (1.5 \times 10^7) (140)$ $= (5.9 \text{ } \mu\text{V})^2$
Amplifier Noise	$\int_0^{140} 3721 en_4^2 \left(\frac{1}{1 + 2\pi \gamma_f f}\right)^2 df = (157 \text{ } \mu\text{V})^2$
	$\bar{V}_{NO} = \left[(12 \text{ } \mu\text{V})^2 + (426 \text{ } \mu\text{V})^2 + (32 \text{ } \mu\text{V})^2 + (5.9 \text{ } \mu\text{V})^2 + (157 \text{ } \mu\text{V})^2 \right]^{1/2} = 455 \text{ } \mu\text{V RMS}$

From an inspection of the output RMS noise equation, it is

apparent that detector noise alone dominates. This would indicate that a lesser grade amplifier could be used with little compromise to the signal-to-noise ratio. However, from a drift standpoint a low offset voltage amplifier must be used.

Let us examine actual data sheet performance on detectors that have been purchased for breadboard testing. Two detectors, .025 x .025 were purchased from Optoelectronics, Inc. They were tested with 12 volts applied across the detector and a 1 M Ω bias resistor. The noise was 2.0 μ V with a 10 Hz bandpass.

$$V_N = \frac{K_N \gamma R_S V_{APP}}{D^*} \sqrt{\ln \frac{600}{590}}$$

The D^* is assumed to be 8×10^9 at 1 Hz

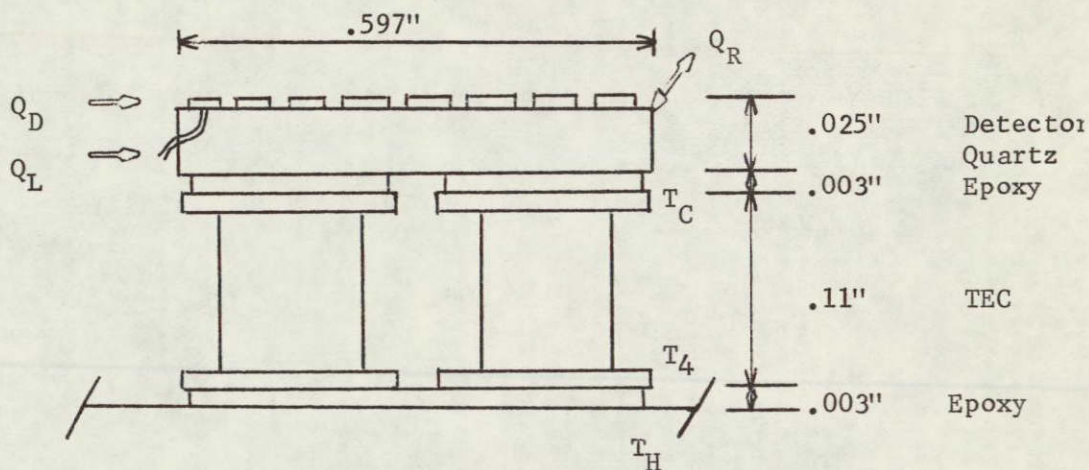
$$V_N = (9.2 \times 10^{-6}) (.1296) = 1.2 \mu V$$

or reasonably close to the tested value. This tends to confirm the fact that the detector noise derivation and D^* extrapolation is correct. This confirmation is made more solid by the tested performance of two sample detector arrays supplied by Optoelectronics. Under the same test conditions described above, the noise ranged from 1.2 μ V to 1.5 μ V. The obvious choice for an amplifier is the LM4250 that was used in the Viking PSA. Specification limits show that power consumption is only 3m watt (for $I_{SET} = 10 \mu A$) and the noise is $50 \text{ NV}/\sqrt{\text{Hz}}$.

APPENDIX C

PSA THERMAL ANALYSIS

Since PbS detectors require tight temperature control, it is necessary to provide a thermoelectric device to cool or heat the detectors. Thermal requirements are that a cooler/heater with dimensions of 0.16" x 0.16" x 0.11" maintain the detector at 25°C over an ambient temperature range of -48 to +63°C. The PbS detector array is deposited on a quartz substrate 0.25" x 0.597" x 0.025", sixteen detector leads 0.1" long of 0.001" gold wire are used, and Epotek H74 thermally conductive epoxy was used. This system is physically characterized as follows:



Thermal conduction in the detector lead wires is described by a thermal conductance, G_L , where

$$G_L = 16 \frac{kA}{\Delta X} = 2.387 \times 10^{-3} = \frac{1}{419}$$

Heat supplied to the detector when the PSA package is at 63°C ("hot case") is

$$Q_L = \frac{1}{419} (145 - 77) = 0.1623 \frac{\text{BTU}}{\text{Hr}} = 0.0476 \text{ watts}$$

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Heat removed with a "cold case" at -48°C is

$$Q_L = \frac{1}{419} \quad 77 - (-55) = 0.315 \frac{\text{BTU}}{\text{Hr}} = 0.0923 \text{ watts}$$

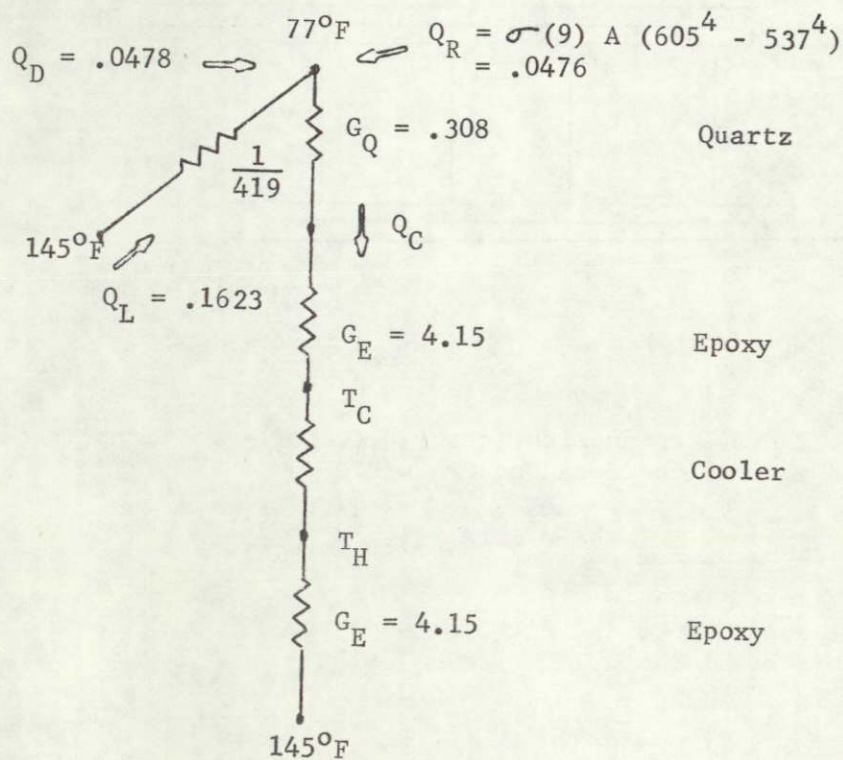
Thermal conduction in the epoxy is described by a thermal conductance, G_E , where

$$G_E = \frac{\frac{1 \times .25 \times .597}{144}}{\frac{.003}{12}} = 4.15$$

Thermal conduction in the quartz substrate is described by a thermal conductance, G_Q , where

$$G_Q = \frac{\frac{.62 \times .25 \times .597}{144}}{\frac{.025}{12}} = 0.31$$

A thermal model for the "hot case" condition is as follows:



Calculation of the cooling required, Q_C , shows

$$Q_C = 0.0478 + 0.0476 + 0.1623 = 0.2577 \frac{\text{BTU}}{\text{Hr}}$$
$$= 0.0755 \text{ watts}$$

Similar calculations for the "cold case" condition show heating required, Q_H , is

$$Q_H = -0.0478 + 0.0527 + 0.315 = 0.3199 \frac{\text{BTU}}{\text{Hr}}$$
$$= 0.0937 \text{ watts}$$

Therefore, the worst-case condition is heating the detector when package is at -48°C . Assuming a 20% efficiency, a thermoelectric device must be capable of handling 0.469 watts. The Marlow Industries MI 1020 single-stage cooler has approximately this capability.

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REFERENCES

1. Huck, Friedrich O.; Katzberg, Stephen J.; and Kelly, W. Lane, IV: The Facsimile Camera--Its Potential as a Planetary Lander Imaging System. Opt. Eng., vol. 11, nos. 4 & 5, July/Oct. 1972, pp. 91-96.
2. Kelly, W. Lane, IV; and Katzberg, Stephen J.: Investigation of Responsivity and Noise in a Direct-Coupled Photodetector-Preamplifier for Facsimile Cameras. NASA TN D-7338, 1973.
3. Kelly, W. Lane, IV; Jobson, Daniel J.; and Rowland, Carroll W.: Design and Evaluation of a Filter Spectrometer Concept for Facsimile Cameras. NASA TN D-7717, 1974.
4. Jobson, Daniel J.; Kelly, W. Lane; and Wall, Stephen D.: A Filter Spectrometer Concept for Facsimile Cameras. NASA TN D-7560, 1974.
5. Kelly, W. Lane, IV: Investigation of Facsimile Camera-Spectrometer Capability in the 1.0 to 2.7 Micron Spectral Range. NASA TM X-72714, 1975.